WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

JANUARY 1952

VOL. XXIX No. 340 THREE SHILLINGS AND SIXPENCE

JANUARY 1952 WIRELESS ENGINEER VOL. XXIX No. 340

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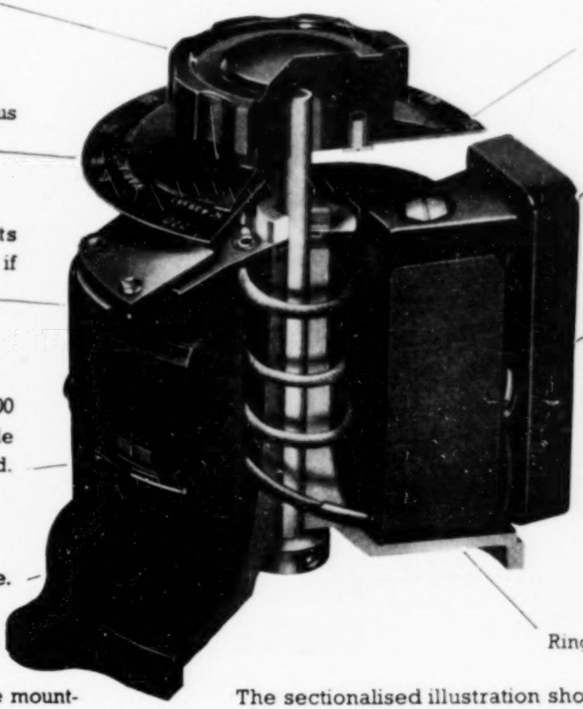

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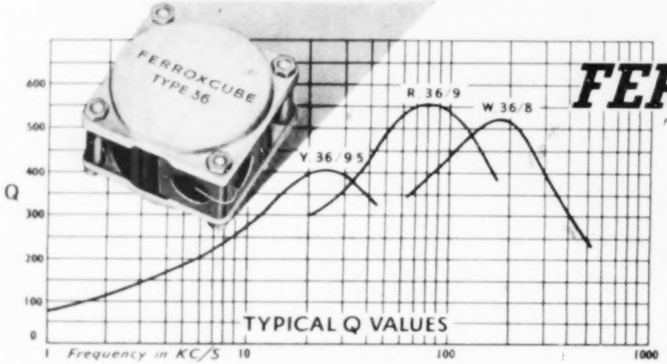

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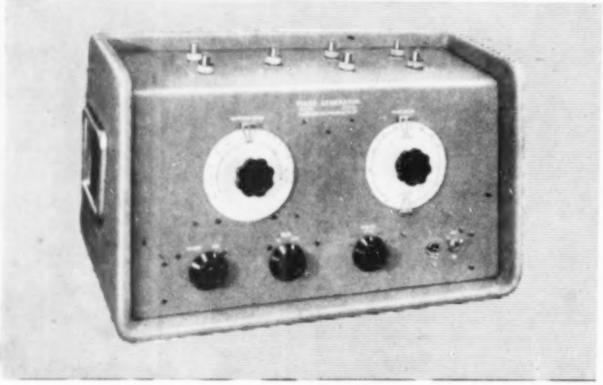

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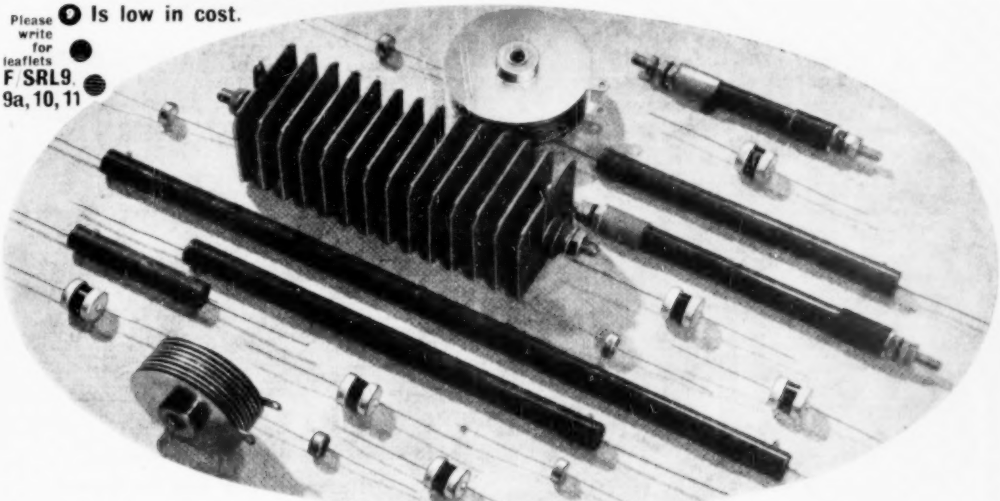

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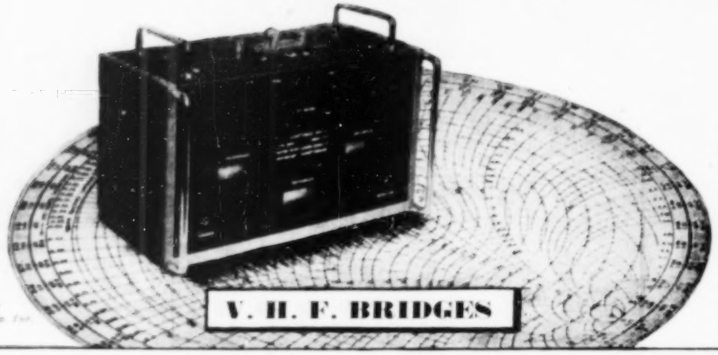
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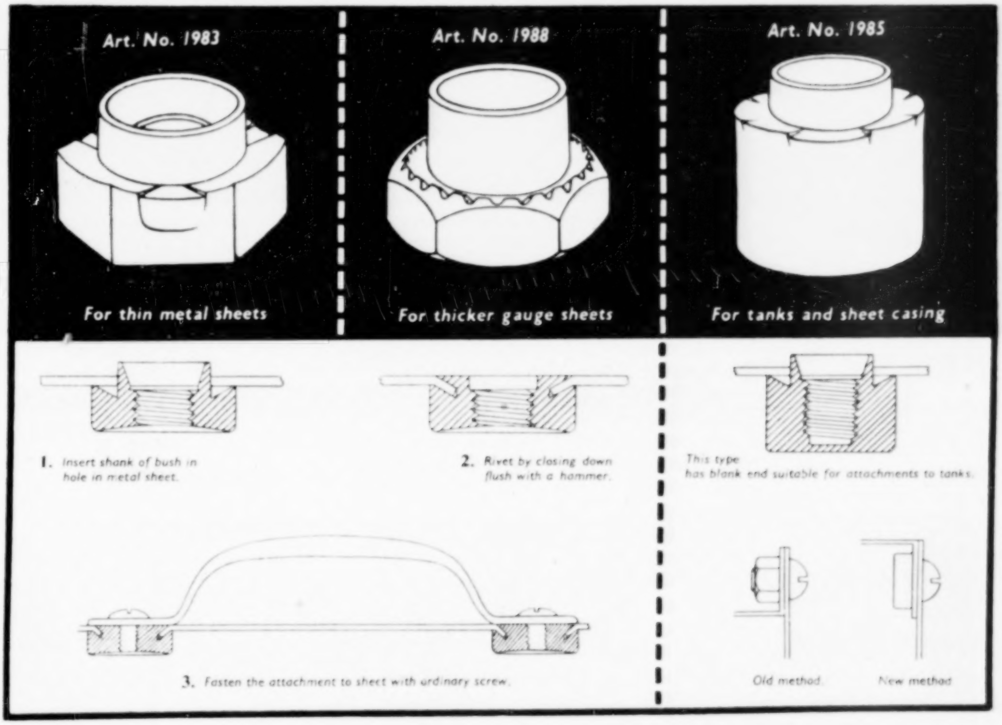

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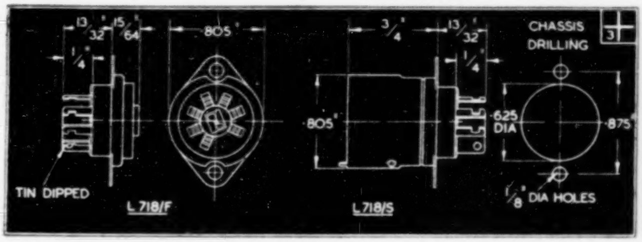
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3

B7G VALVEHOLDER

LIST NUMBERS

L718 F	Flange
*Z560132	
L718 S	Skirt
*Z560127	
L719	Screening Can

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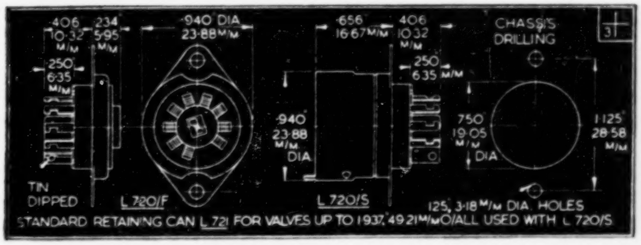
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·250 6·35 M/M
·940 DIA 23·88 M/M
TIN DIPPED
L 720/F
·656 16·67 M/M
·406 10·32 M/M
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·940 23·88 M/M DIA
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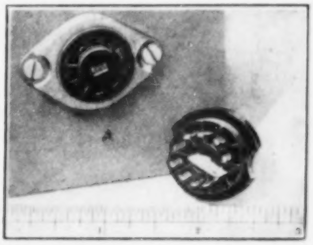

B9A "NOVAL" VALVEHOLDER

LIST NUMBERS

L720 F	Flange
L720 S	Skirt
L721	Screening Can

The "Belling-Lee" version of this valveholder is now in its initial production stage and samples are awaiting the approval of the appropriate authority.

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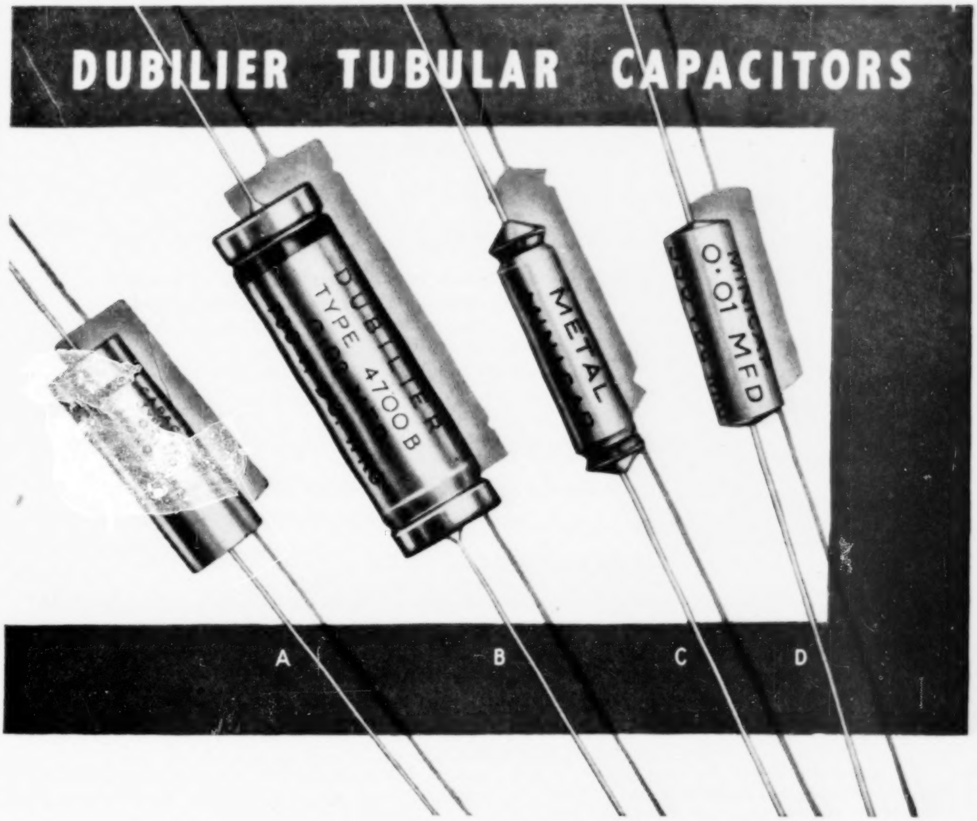

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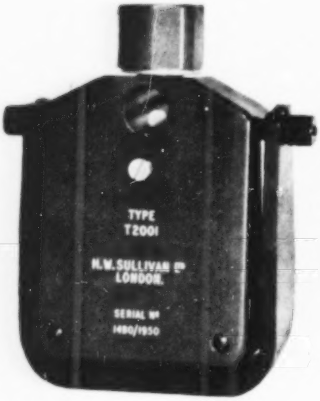

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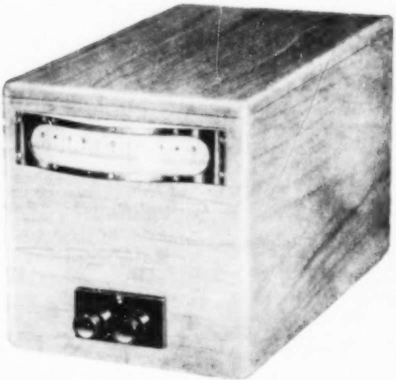

WIRELESS ENGINEER

The Journal of Radio Research and Progress

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JANUARY 1952

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ABSTRACTS AND REFERENCES. Nos. 1-292 **A.1-A.22**

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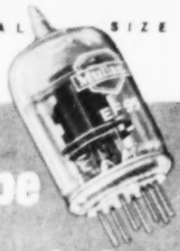
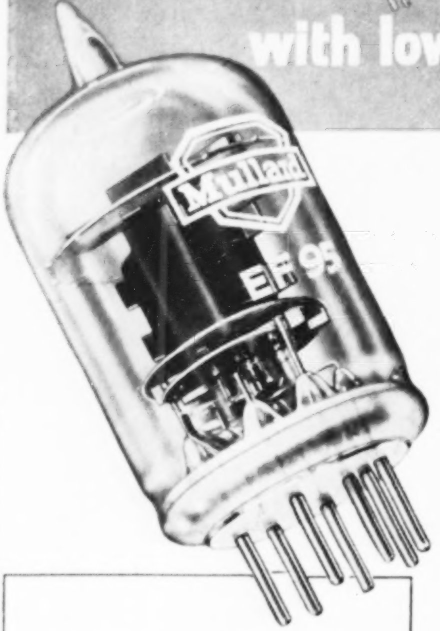


A High-Slope Low-Noise R.F. Pentode with low power consumption



The combination of good slope to capacity ratio, low noise factor, and low power consumption makes this new Mullard R.F. pentode an ideal valve for use in all types of communications equipment operating at V.H.F. Its slope to capacity ratio approaches unity, and this makes it specially suitable for use in Wide Band Amplifiers.

Used as a neutralized triode, and followed by a grounded grid stage (cascode), the EF95 has a typical noise factor of 3.5 when operated at 180 Mc/s.

With a heater current of only 0·175A., this valve shows a large saving in power consumption over previous valves of a similar class. This saving in power is emphasized in applications where large numbers of R.F. stages are involved.

The EF95 has identical characteristics to the American 6AK5 and may be used as a direct replacement for this valve.

For full technical information on this and other valves in the Mullard range please write to the address below.

Heater		**Capacitances**	
V_h	6.3 V	c_{in}	4.0 μμF
I_h	0.175 A	c_{out}	2.8 μμF
		c_{a-g1}	0.02 μμF
Characteristics			
V_a	120 V		
V_{g2}	120 V	**Limiting Values**	
V_{g1}	−2.0 V	V_a max.	180 V
I_a	7.5 mA	p_a max.	1.7 W
I_{g2}	2.5 mA	V_{g2} max.	140 V
g_m	5.0 mA/V	p_{g2} max.	0.5 W
r_a	0.34 MΩ	I_k max.	18 mA

Base B7G

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Vol. XXIX JANUARY 1952 No. 340

Cavity Resonators

IN the early days of radio the currents flowed on the outer surfaces of the conductors and the electric and magnetic fields spread out into the surrounding space. In recent years, with the development of microwaves both in transmission and research, the tendency has been in the opposite direction, with the currents flowing on the inner surfaces of hollow conductors and the electric and magnetic fields confined entirely to the internal space. We have been struck by the growing use of the cavity resonator as a research device. In the April Editorial we referred to its use at the N.P.L. and at Stanford University for determining the velocity of electromagnetic waves in a vacuum. This necessitates the very accurate determination of the frequency and of the dimensions of the cavity; as we said in April, it is very satisfactory that the results obtained at the N.P.L. and Stanford, and those obtained by radar and optical methods, all agree within 4 parts in 300,000.

Much research work with cavity resonators has been carried out in the Research Laboratories of the Philips Company. An interesting application was the determination of the dielectric constant and loss angle of various dielectrics at a frequency of 3,000 Mc/s; it was stated that if the wrong type of glass is used for the bulb of a transmitting valve, it may melt owing to excessive dielectric losses.[1] As the loss angle may vary erratically with the frequency, it is important to make measurements at the actual frequency at which the material is to be used.

The method employed is the well-known one of taking a resonant circuit and finding the effect of introducing the dielectric into the gap of the capacitor. From the change in the resonant frequency and the broadening of the resonance curve one can determine both the dielectric constant ϵ and the loss angle δ. With lumped inductance and capacitance at a low frequency this is a simple matter, but it becomes more complicated with a cavity resonator at a frequency of 3,000 Mc/s.

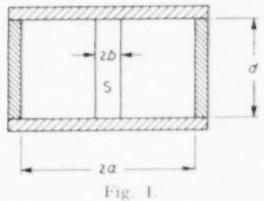

Fig. 1.

Fig. 1 shows the type of cavity employed; the cylinder of diameter $2a$ and height d has a top and a bottom plate, between which stands the rod S of the dielectric under investigation; the actual cylinder was about 3 in. diameter and $1\frac{3}{4}$ in. high. For accurate measurement of the dielectric loss in the specimen it is essential that the losses in the cavity itself should be as small as possible; this necessitates very good contact between the cylinder and the top and bottom plates. The fact that the measured loss was always greater than the calculated value was found to be due to the corrugated nature of the inner surface of the cylinder when turned in the ordinary way. Since the depth of penetration is only about 1μ, the actual length of the current path and therefore the loss, is increased by the corrugation. To reduce this to a minimum the silver-plated cylinder was turned in a precision lathe with a diamond of semi-circular profile, the pitch being about 1μ and the shaving thickness extremely small. By these means the Q value of the empty resonator was

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[1] See article by M. Gevers in *Philips Technical Review*, September 1951, p. 61, where many other references are given.

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raised to 98 per cent of the theoretical value. The dielectric rod must also be prepared with care, the surface being smooth and the diameter the same throughout. Apparently its exact position is not so important and it can be centred by eye.

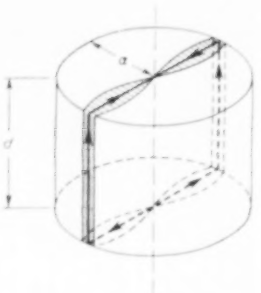


Fig. 2 (left).

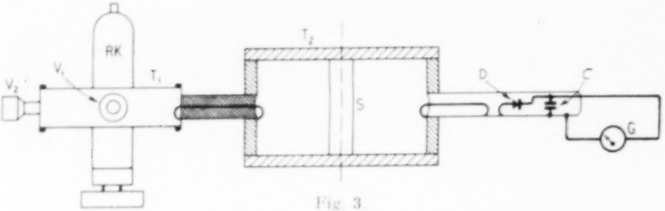


Fig. 3.

From Fig. 2 it is evident that the cavity resonator may be regarded as a large number of half-wave oscillators of length $d + 2a$, bent over at right-angles at the top and bottom and connected in parallel. The width of the shaded strip shows the current density at the fundamental mode; it is approximately constant along the side and reaches a maximum at a point in the top and bottom where the increasing current is counterbalanced by the increasing cross-section. The oscillations are produced by means of a reflex klystron RK (Fig. 3) coupled to the cavity T_2 by means of loops and short lengths of coaxial cable; in the same way the cavity is coupled to a silicon-crystal detector D which is connected to a smoothing capacitor C and the galvanometer G. The coaxial cable between the cavity T_1 of the oscillator and the cavity T_2 containing the test specimen S is filled with a material of high dielectric loss to prevent resonances occurring in it. Similar precautions are adopted in the connection between T_2 and the detector. The frequency of the klystron oscillator can be adjusted by the micrometer screws V_1 and V_2, and read off calibration curves. Measurements are made of the resonant frequency and of the frequency variations necessary to reduce the deflection to $1/\sqrt{2}$ of the resonant value; these are made first with the empty cavity and then with the dielectric rod in position. The results enable ϵ and $\tan\delta$ of the dielectric to be calculated. The following three examples show how the losses may or may not vary with the frequency.

Frequency in Mc/s	3,100	1·5
	$\tan\delta \times 10^4$	$\tan\delta \times 10^4$
A hard glass	64·8	36·4
Quartz glass	0·8	0·8
Perspex	64·1	130

The presence of the rod affects the distribution of the field to an extent depending on its diameter and its dielectric properties. Relatively simple formulae can be used to give a good approximation if the ratio a/b exceeds $5\sqrt{\epsilon_r}$, where a and b are the cavity and rod radii and ϵ_r is the relative dielectric constant; as ϵ_r increases the radius b of the rod must be made smaller. It is stated that a material such as barium titanate, which may have an ϵ_r of 1,000, would have to be made in the form of a wire of less than 1 mm diameter and passed through holes in the top and bottom plates. The mathematical development of the necessary formulae is given in an appendix to the original article.

G. W. O. H.

ABSTRACTS AND REFERENCES INDEX

The Index to the Abstracts and References published throughout 1951 is in course of preparation and will, it is hoped, be available in February, price 2s. 8d. (including postage). As supplies are limited our Publishers ask us to stress the need for early application for copies. Included with the Index is a selected list of journals scanned for abstracting, with publishers' addresses.

FOURIER ANALYSIS AND NEGATIVE FREQUENCIES

By I. J. Shaw, Ph.D.

1. Introduction

It is well known that a fluctuation $f(t)$, (e.g., of voltage with time) may, by means of Fourier analysis, be represented as the sum of an infinite series of sine and cosine waves of appropriate amplitudes. This series is called the frequency spectrum of the fluctuation and is denoted by $s(\omega)$, where $s(\omega)$ represents the amplitude and phase of the component of angular frequency ω.

The mathematical equations giving the relationship between $f(t)$ and $s(\omega)$ are generally written in the complex algebraic form:

$$\left.\begin{aligned} s(\omega) &= \int_{-\infty}^{\infty} f(t) \exp(-j\omega t)\, dt \\ f(t) &= \frac{1}{2\pi}\int_{-\infty}^{\infty} s(\omega) \exp(j\omega t)\, d\omega \end{aligned}\right\} \quad \ldots \quad (1)$$

The exponential terms in these equations may be regarded as vectors of unit amplitude rotating with an angular velocity of ω radians per second. The term $\exp(-j\omega t)$ represents a vector rotating in the clockwise direction, and $\exp(j\omega t)$ one rotating in the anti-clockwise direction. From the second equation of the pair, we see, for example, that to build up the fluctuation $f(t)$ we must find the sum of a series of vectors, the amplitude and phases of which are given by the appropriate value of $s(\omega)$, for the angular frequency ω which is summed over the range from $-\infty$ to $+\infty$.

So long as we are content to regard $s(\omega)$ as giving the amplitude and phase of a vector rotating with an angular frequency ω, the presence of negative frequencies seems more or less natural. But for practical purposes we require the amplitude and phase of a series of sinusoidal waves and then the idea of negative frequencies becomes somewhat less tangible. One thinks perhaps, that, like the negative square root of a positive number, negative frequencies may be of mathematical interest but of little practical importance.

The work described in this paper is the result of just such an approach to the problem. In an application of the 'modulation theorem' (to be described later) in Fourier analysis, it was found that correct results were obtained in many cases, but that in certain cases the spectra found by this means were not the same as those obtained by direct application of equation (1). It was then discovered that this error arose through the neglect of a frequency spectrum centred on a negative frequency.

Incidental to the investigation of the effects of negative frequencies two comparatively easy methods have been found for obtaining the frequency spectrum of a block of sine or cosine waves of a given number of cycles duration and given initial phase. These may commend themselves to those more at home with graphical methods or with straightforward trigonometrical formulae than with the integration and complex algebra involved in the evaluation of equation (1).

In a later section we extend our investigation to a consideration of aerial polar diagrams which can be deduced from the Fourier analysis of the distribution of current across the aerial aperture. Here we shall find the idea of negative frequencies replaced by the more familiar concept of negative angles. Aerial polar diagrams may, in many instances, be determined easily from first principles, and while it is not suggested that in such cases the approach by way of Fourier analysis is any simpler, it may be found interesting because it gives us a method of seeing how the polar diagram is made up. As will be seen later the physical understanding gained by the Fourier-analysis approach enables us to see in what way an aerial must be modified to effect a desired change in its characteristics, shows us where approximations may safely be made, and helps to develop the relationship between resonant and non-resonant aerials.

2. The Modulation Theorem

The modulation theorem of Fourier analysis is generally well known. Stated mathematically it is:—

> "If a fluctuation $f(t)$ has a frequency spectrum $s(\omega)$, then the frequency spectrum of the modulated fluctuation $f(t)\exp(j\omega_c t)$ is $s(\omega - \omega_c)$."

The theorem is easily proved thus:—

By definition the frequency spectrum $s(\omega)$ of the fluctuation $f(t)$ is given by

$$s(\omega) = \int_{-\infty}^{\infty} f(t) \exp(-j\omega t)\, dt$$

Let $s'(\omega)$ be the frequency spectrum of $f(t)\exp(j\omega_c t)$.

MS accepted by the Editor, November 1950

Then

$$s'(\omega) = \int_{-\infty}^{\infty} f(t) \exp(j\omega_c t) \exp(-j\omega t)\, dt$$

$$= \int_{-\infty}^{\infty} f(t) \exp\left[-j(\omega - \omega_c)\, t\right] dt$$

$$= s(\omega - \omega_c) \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

Thus, if we have the spectrum of the fluctuation $f(t)$, then the spectrum of the modulated fluctuation $f(t) \exp(j\omega_c t)$ is precisely the same except that it is now centred on the frequency ω_c instead of on zero frequency.

3. Examples of Use of Modulation Theorem

In our investigation we shall find that the fluctuation $f(t)$ which is most useful is the unit pulse and we shall need to know its frequency spectrum $s(\omega)$. This is commonly known, but we shall derive it here for the sake of completeness.

The unit pulse is a function which is of unit amplitude for the duration of the pulse and zero everywhere else. For a reason which will be obvious later, we shall assume that the pulse of duration T begins at time $t = -T/2$ and ends at $t = T/2$. Then the frequency spectrum is given by

$$s(\omega) = \int_{-\infty}^{\infty} f(t) \exp(-j\omega t)\, dt$$

$$= \int_{-T/2}^{T/2} \exp(-j\omega t)\, dt$$

$$= \frac{-1}{j\omega}\left[\exp(-j\omega T/2) - \exp(j\omega T/2)\right]$$

$$= \frac{2 \sin \omega T/2}{\omega} \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

Thus with a pulse of unit amplitude and duration T centred on $t = 0$, the frequency spectrum is given by equation (3) and we notice in particular that equation (3) is a function which is entirely real. This means that the frequency components are all cosine waves in phase or in antiphase at $t = 0$. This is a simplicity which we wish to preserve, and it was for this reason that the pulse was centred on $t = 0$.

We shall now define the frequency ω_0 such that T is the time of one cycle.

i.e., $T = 2\pi/\omega_0$

$\therefore \ \omega_0 = 2\pi/T$

The frequency spectrum of the pulse may now be written

$$s(\omega) = \frac{2 \sin \omega T/2}{\omega} = \frac{2 \sin \omega\pi/\omega_0}{\omega} \quad \ldots \quad (4)$$

It is easy to see that $s(\omega) = 0$ for $\omega = n\omega_0$ where n is any integer except zero. For $\omega = 0$, the function may be evaluated as

$$s(\omega) = \frac{2\pi}{\omega_0} \cdot \frac{\sin \frac{\omega}{\omega_0}\pi}{\frac{\omega}{\omega_0}\pi}$$

The value of $\frac{\sin \theta}{\theta}$ for $\theta \rightarrow 0$ is unity, and we have

therefore $\qquad s(o) = \frac{2\pi}{\omega_0}$

Table 1 gives some of the values of $s(\omega)$ arranged in three columns the more clearly to illustrate the symmetry of the function, while a graph of the function is shown in Fig. 1.

Suppose now that we wish to determine the frequency spectrum of a pulse modulated by a cosine wave as shown in Fig. 2, where there is a large number of cycles of the cosine wave within the length of the pulse. [We should more normally call this a cosine wave amplitude modulated 100% by a rectangular pulse. But we shall retain this 'inverted' nomenclature here since the pulse is our original waveform, and we are

TABLE 1.

ω/ω_o	$s(\omega) = \dfrac{2 \sin \frac{\omega\pi}{\omega_o}}{\omega}$		
0	$\frac{2\pi}{\omega_o}$		
0·25		$\frac{4\sqrt{2}}{\omega_o}$	
0·5			$\frac{4}{\omega_o}$
0·75		$\frac{4\sqrt{2}}{3\,\omega_o}$	
1·0	0		
1·25		$\frac{-4\sqrt{2}}{5\,\omega_o}$	
1·5			$\frac{-4}{3\,\omega_o}$
1·75		$\frac{-4\sqrt{2}}{7\,\omega_o}$	
2·0	0		
2·25		$\frac{4\sqrt{2}}{9\omega_o}$	
2·5			$\frac{4}{5\omega_o}$
2·75		$\frac{4\sqrt{2}}{11\omega_o}$	
3·0	0		
3·25		$\frac{-4\sqrt{2}}{13\omega_o}$	
3·5			$\frac{-4}{7\omega_o}$
3·75		$\frac{-4\sqrt{2}}{15\omega_o}$	
4·0	0		

about to relate the modulating vector exp $(j\omega_c t)$ of the modulation theorem to the cosine wave contained within the pulse.]

In the modulation theorem which was quoted above, the pulse was modulated by the rotating unit vector exp $(j\omega_c t)$. In order to transform this rather mathematical concept to the real case of a pulse modulated by a cosine wave, we need merely use the equation

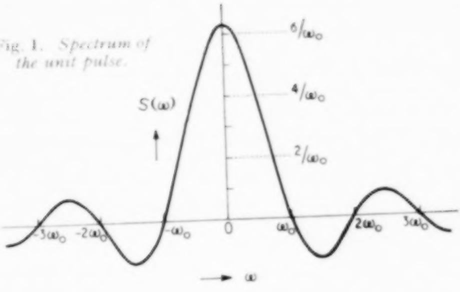


Fig. 1. *Spectrum of the unit pulse.*

$$\cos \omega_c t = \tfrac{1}{2}\left[\exp(j\omega_c t) + \exp(-j\omega_c t)\right] \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

This relation states mathematically that a unit cosine function can be represented as half the sum of two unit vectors rotating in opposite directions with an angular frequency equal to that of the cosine function.

The frequency spectrum $s'(\omega)$ of the cosine modulated pulse is

$$s'(\omega) = \int_{-\infty}^{\infty} f(t) \cos \omega_c t \exp(-j\omega t)\, dt$$

$$= \tfrac{1}{2} \int_{-T/2}^{T/2} \left[\exp(j\omega_c t) + \exp(-j\omega_c t)\right] \exp(-j\omega t)\, dt$$

$$= \tfrac{1}{2} \int_{-T/2}^{T/2} \left[\exp\{-j(\omega - \omega_c)t\} + \exp\{-j(\omega + \omega_c)t\}\right] dt$$

$$= \tfrac{1}{2}\left[s(\omega - \omega_c) + s(\omega + \omega_c)\right] \quad \ldots \quad (6)$$

where $s(\omega)$ is the frequency spectrum of the unit pulse of duration T centred on time $t = 0$. We thus see that in order to find the frequency spectrum of the modulated pulse we need merely to take half the sum of two spectra, one of which is the spectrum of the pulse centred on ω_c, and the other that of the pulse centred on $-\omega_c$. We have assumed here that there are many cycles of cosine oscillation within the pulse and thus $\omega_c \gg \omega_0$. The two spectra which we have to add are shown in Fig. 3 and if, in a practical way, we consider the amplitude of the frequency components in the positive half of the frequency plane, we see that contributions from the spectrum centred on $-\omega_c$ are so small in the positive portion of the diagram that we may safely disregard them. The frequency spectrum of the modulated pulse is then just the frequency spectrum of the pulse itself centred on the frequency ω_c. In this case we have, in fact found the spectrum of a section of a cosine wave containing many cycles of oscillation.

Suppose now that instead of a section of a cosine wave containing many cycles, we wish to find the spectrum of a fluctuation which is only one half of a single cycle of a cosine wave, as shown in Fig. 4.

In the same way as before we shall consider that this fluctuation is, in fact, a pulse modulated by a cosine wave. The treatment will be exactly the same as in the previous case, but since the pulse is our primary waveform we shall assume, not that the pulse is shorter than before, but that the frequency of the cosine wave is less than before. If T_c is the period of the cosine wave, and ω_c its frequency, we have

$$T_c = \frac{2\pi}{\omega_c} = 2T = 2\,\frac{2\pi}{\omega_0}$$

$$\therefore\ \omega_c = \tfrac{1}{2}\,\omega_0$$

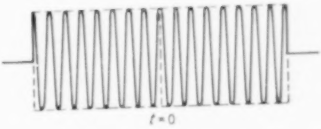


Fig. 2 (above). *Pulse modulated by a cosine wave.*

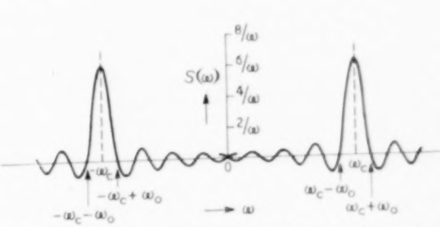


Fig. 3 (below). *Component spectra of the pulse of Fig. 2.*

To find the frequency spectrum we must now take half the sum of the two spectra as shown in Fig. 5, centred on $\pm\,\omega_c$; i.e., on $\pm\,\tfrac{1}{2}\omega_0$. In this case, the spectrum centred on $-\omega_c$ spreads well into the positive half of the frequency plane and, in calculating the resultant, contributions from it may not be neglected as they were in the previous case. The final spectrum is found by taking half the sum of the component spectra, taking note of the fact that portions of the spectra appearing below the line must be regarded as having negative amplitude. Fig. 6 shows the spectrum obtained.

We now have a method for determining rapidly the frequency spectrum of a portion of a cosine wave containing any number of cycles of oscillation, provided that the portion is symmetrically placed about the $t = 0$ axis. All we have to do is to draw on transparent paper two identical patterns representing the spectrum of a pulse (Fig. 1), superimpose them, and displace by the required amount, centring the spectra on $\pm\ \omega_c$, and add the overlapping portions in the positive half of the frequency plane.

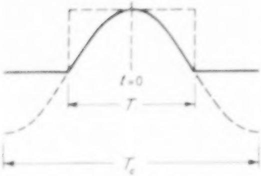


Fig. 4 (above). *Pulse comprising one half-cycle of a cosine wave.*

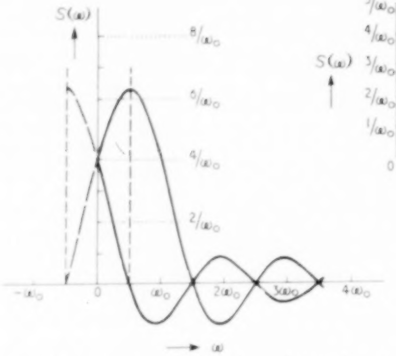


Fig. 5 (right). *Component spectra of the pulse of Fig. 4.*

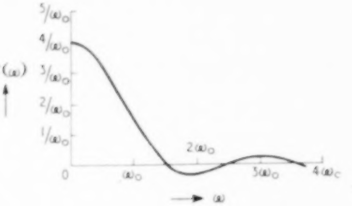


Fig. 6 (above). *Sum of the spectra of Fig. 5.*

In many cases, however, we shall be interested in the frequency spectrum of a block of oscillations which is not symmetrically placed inside the enveloping pulse. Let us consider, for example, the frequency spectrum of the fluctuation shown in Fig. 7, which is a section of a cosine wave of half a cycle duration, but which is not symmetrical about time $t = 0$. It is obvious from physical considerations that the frequency spectrum of this will not be the same as that of the fluctuation shown in Fig. 4 since now we have two sharp edges to deal with which were previously absent and, furthermore, the d.c. components of the two fluctuations are obviously different.

At this point, it might be asked why it was, when considering the frequency spectrum of the block of a large number of oscillations (Fig. 2) that this question of the phase of the oscillations did not arise, and that the frequency spectrum of the fluctuation which we obtained was independent of whether the oscillations began from unit amplitude (cosine wave), zero amplitude (sine wave) or from some intermediate point. The answer is that, in fact, the initial phase does matter and comes into the final answer, as we shall see later, by determining the phase relation between the two spectra which we add together. But in this particular case, the contributions from the spectrum centred on $-\omega_c$ were so small that they could be disregarded. From the physical point of view we may see that in this case of many oscillations, the initial phase will have only a very small effect on the resultant spectrum from the fact that, since there are so many cycles in the pulse, the rate of rise of the leading edge is much the same whether we change abruptly from zero to unit amplitude at this point, or whether we approach unit amplitude from zero following a sinusoidal curve.

Returning now to the fluctuation of Fig. 7, at the centre of the pulse (i.e., at time $t = 0$), the cosine function has the value $\cos\phi$, and hence the total cosine function must be written as $\cos(\omega_c t + \phi)$. The cosine function may therefore be expressed in the exponential form as

$$\cos(\omega_c t + \phi) = \tfrac{1}{2}\left[\exp\{j(\omega_c t + \phi)\} + \exp\{-j(\omega_c t + \phi)\}\right]$$

and the frequency spectrum $s'(\omega)$ of $f(t)\cos(\omega_c t + \phi)$ becomes

$$\begin{aligned} s'(\omega) &= \tfrac{1}{2}\int_{-T/2}^{T/2}\left[\exp\{j(\omega_c t + \phi)\} + \exp\{-j(\omega_c t + \phi)\}\right]\exp(-j\omega t)\,dt \\ &= \tfrac{1}{2}\int_{-T/2}^{T/2}\left[\exp\{-j(\omega - \omega_c)t\}\exp(j\phi) + \exp\{-j(\omega + \omega_c)t\}\exp(-j\phi)\right]dt \\ &= \tfrac{1}{2}\left[s(\omega - \omega_c)\exp(j\phi) + s(\omega + \omega_c)\exp(-j\phi)\right] \quad \cdots \quad (7) \end{aligned}$$

where $s(\omega)$ is the frequency spectrum of the unit pulse $f(t)$.

This equation tells us that the frequency spectrum of the modulated pulse is equal, as before, to half the sum of the two displaced spectra, but that this time, the phase of each component of the spectrum centred on frequency ω_c is advanced by an angle ϕ while the phase of each component of the spectrum centred on $-\omega_c$ is retarded by the same angle ϕ, before the addition is carried out.

For the fluctuation shown in Fig. 7 the phase angle ϕ is 45°. Since the frequency ω_c of the cosine wave bears the same relation to ω_0 of the pulse as in the previous case of Fig. 4, the component spectra are positioned exactly as previously (Fig. 5). In obtaining the final spectrum, however, we must treat the various amplitudes as vectors and advance the phases of those centred on ω_c by 45° and retard the phases of those from the spectrum centred on $-\omega_c$ by 45° before adding. Certain points on the final frequency spectrum are immediately obvious; e.g., at all points where the amplitude of one of the component spectra is zero, the resultant is equal to one-half the value of the corresponding component of the remaining spectrum. A few other points are easily calculated and the final spectrum of Fig. 8 is soon obtained.

In Fig. 8, the spectrum is drawn entirely in the positive quadrant of the diagram because the phases of the various component frequencies are all different. In Fig. 6, however, the phases of the components were all zero or 180°, and this fact is conveniently represented by using negative amplitudes as shown.

The process of combining the amplitudes of the overlapping spectra may, if desired, be carried out by calculation rather than by graphical means.

The frequency spectrum of the pulse centred on ω_c is

$$\frac{2\sin\left(\dfrac{\omega-\omega_c}{\omega_0}\right)\pi}{(\omega-\omega_c)} \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

and that of the pulse centred on $-\omega_c$ is

$$\frac{2\sin\left(\dfrac{\omega+\omega_c}{\omega_0}\right)\pi}{(\omega+\omega_c)} \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

Write $\omega_c = x\omega_0$ and the spectra become

$$\frac{2}{\omega_0}\cdot\frac{\sin\left(\dfrac{\omega}{\omega_0}-x\right)\pi}{\left(\dfrac{\omega}{\omega_0}-x\right)} \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

and

$$\frac{2}{\omega_0}\cdot\frac{\sin\left(\dfrac{\omega}{\omega_0}+x\right)\pi}{\left(\dfrac{\omega}{\omega_0}+x\right)} \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

At any frequency ω, the corresponding values of equations (10) and (11) must be added with a phase difference of 2ϕ, and one-half of this resultant will give the amplitude of the component of frequency ω in the final spectrum.

Thus we have:—

$$2\,s'(\omega) = \left[\frac{4}{\omega_0^2}\cdot\frac{\sin^2\left(\dfrac{\omega}{\omega_0}-x\right)\pi}{\left(\dfrac{\omega}{\omega_0}-x\right)^2} + \frac{4}{\omega_0^2}\cdot\frac{\sin^2\left(\dfrac{\omega}{\omega_0}+x\right)\pi}{\left(\dfrac{\omega}{\omega_0}+x\right)^2}\right.$$
$$\left. + 2\cdot\frac{2}{\omega_0}\,\frac{\sin\left(\dfrac{\omega}{\omega_0}-x\right)\pi}{\left(\dfrac{\omega}{\omega_0}-x\right)}\cdot\frac{2}{\omega_0}\cdot\frac{\sin\left(\dfrac{\omega}{\omega_0}+x\right)\pi}{\left(\dfrac{\omega}{\omega_0}+x\right)}\cdot\cos 2\phi\right]^{\frac{1}{2}}$$

which, on simplification, can be shown to reduce to

$$s'(\omega) = \frac{1}{\omega_0}\left[\left\{\frac{\sin\left(\dfrac{\omega}{\omega_0}-x\right)\pi}{\left(\dfrac{\omega}{\omega_0}-x\right)} + \frac{\sin\left(\dfrac{\omega}{\omega_0}+x\right)\pi}{\left(\dfrac{\omega}{\omega_0}+x\right)}\right\}^2\right.$$
$$\left. - 4\,\frac{\sin\left(\dfrac{\omega}{\omega_0}-x\right)\pi}{\left(\dfrac{\omega}{\omega_0}-x\right)}\cdot\frac{\sin\left(\dfrac{\omega}{\omega_0}+x\right)\pi}{\left(\dfrac{\omega}{\omega_0}+x\right)}\cdot\sin^2\phi\right]^{\frac{1}{2}} \qquad (12)$$

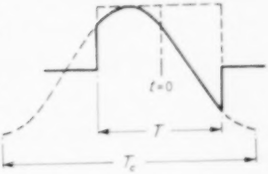


Fig. 7 (right). *Unsymmetrical pulse of one half-cycle duration.*

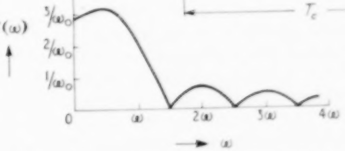


Fig. 8 (below). *Spectrum of pulse of Fig. 7.*

Although equation (12) may look complicated, it is quite easy to use in calculation, and the process of integration and obtaining the modulus of equation (1) has been replaced by direct substitution into a trigonometrical formula. The only troublesome point is met when the substitution $\omega/\omega_0 = x$ is made, but this is easily overcome by remembering that the limiting value of $\frac{\sin\theta}{\theta}$ as $\theta \rightarrow 0$ is unity.

As an example of the method we shall determine how the frequency spectrum of a section of a cosine wave, of one half-cycle duration, varies as the phase is altered.

If T_o is the length of the enveloping pulse we have

$$T_c = 2T_o$$

$$\therefore \frac{2\pi}{\omega_c} = 2 \cdot \frac{2\pi}{\omega_o}$$

$$\text{whence } \omega_c = \tfrac{1}{2}\,\omega_o$$

$$\text{and } x = \tfrac{1}{2} \cdot$$

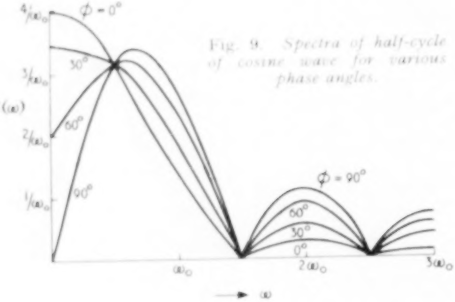


Fig. 9. *Spectra of half-cycle of cosine wave for various phase angles.*

We now evaluate (12) with $x = \frac{1}{2}$ for different values of ϕ. The result is shown in Fig. 9. We notice, as is to be expected, that as the function changes from a half-cycle with a maximum at $t = 0$ ($\phi = 0$, half-cycle of cosine wave) to a half-cycle with zero at $t = 0$ ($\phi = 90°$, half-cycle sine wave), the d.c. component becomes progressively less and the amplitude of the higher frequency components progressively greater.

4. Application to the Calculation of Aerial Polar Diagrams.

Suppose we have an aerial of length L with a distribution of current along it given by

$$I = I_o f(x)$$

where x is the distance of the point under consideration from the mid-point ($x = 0$) of the aerial, and we wish to calculate the field at a point P as shown in Fig. 10. This calculation is made by adding, in the correct phase, all the small fields radiated from the elementary doublets of which we may consider the aerial to be made. We know that the field at P due to an elementary doublet is proportional to the length of the doublet, the current flowing in it and the cosine of the angle θ. Considering the point R on the aerial we notice that the field from the length dx at R will lead the field from the length dx at the origin by an angle ϕ, where

$$\phi = \frac{2\pi}{\lambda} ON = \frac{2\pi x}{\lambda} \sin\theta \quad \ldots \quad \ldots \quad (13)$$

where λ is the wavelength of the radiation from the aerial.

The magnitude of the field at P is the vector sum of all such elementary fields each with its appropriate phase, and using the well-known formula for vector addition we have

$$|F_p| = \left[\left\{ \sum_{x=-L/2}^{L/2} F_e \cos\phi_e \right\}^2 + \left\{ \sum_{x=-L/2}^{L/2} F_e \sin\phi_e \right\}^2 \right]^{\frac{1}{2}} \quad \ldots \quad (14)$$

where F_e is the elementary field due to the current in the element dx at x, and ϕ_e is the corresponding phase angle.

For the purposes of calculation this may be written more concisely:

$$F_p = \sum_{x=-L/2}^{L/2} F_e (\cos\phi_e + j\sin\phi_e)$$

which in the integral form becomes

$$F_p = \int_{L/2}^{L/2} RI_o f(x) \cos\theta \exp(j\phi)\, dx$$

$$= RI_o \cos\theta \int_{-\infty}^{\infty} f(x) \exp(jk\,Sx)\, dx \quad \ldots \quad (15)$$

where $k = \frac{2\pi}{\lambda}$, $S = \sin\theta$, R is a constant, and I_o is the current at some reference point. The change of the limits of integration from $\pm L/2$ to $\pm \infty$ does not affect the result, since the aperture distribution, $I_o f(x)$, of current is zero outside $x = \pm L/2$, and this change is made merely to emphasise a similarity which we shall shortly be considering.

If we take P at a fixed distance from the aerial, then F_p, the field at P, is a function of $\sin\theta$ and we may write

$$P(S) = RI_o \cos\theta \int_{-\infty}^{\infty} f(x) \exp(j\,k\,S\,x)\, dx \quad \ldots \quad (16)$$

where $P(S)$, being the field-strength of the wave radiated in a direction defined by $S = \sin\theta$ may be regarded as the polar diagram of the aerial.

If we compare equation (16)

$$P(S) = RI_o \cos\theta \int_{-\infty}^{\infty} f(x) \exp(j\,k\,S\,x)\, dx$$

with equation (1)

$$s(\omega) = \int_{-\infty}^{\infty} f(t) \exp(-j\omega t)\, dt$$

we notice that, apart from constants and the $\cos\theta$ term outside the integral in (16), the two equations are formally equivalent. This suggests that the calculation of polar diagrams of aerials of given aperture distribution is closely analogous to the calculation of frequency spectra of fluctuations of a given form. This fact is indeed well known. What is, perhaps, not quite so well known is that the modulation theorem which we found to be so useful in the calculation of frequency spectra has a corresponding application in the calculation of aerial polar diagrams.

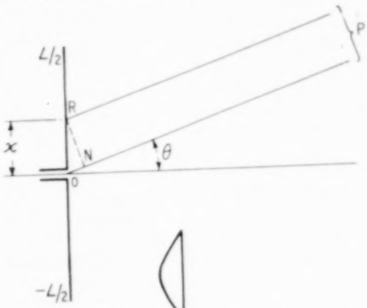


Fig. 10 (above). *Diagram of an aerial, showing the co-ordinates to P.*

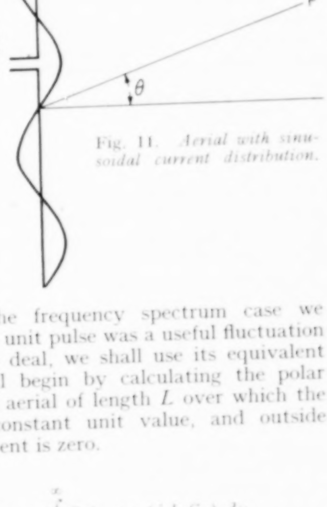


Fig. 11. *Aerial with sinusoidal current distribution.*

Since, in the frequency spectrum case we found that the unit pulse was a useful fluctuation with which to deal, we shall use its equivalent here and shall begin by calculating the polar diagram of an aerial of length L over which the current has constant unit value, and outside which the current is zero.

Then

$$P(S) = RI_o \cos\theta \int_{-\infty}^{\infty} f(x) \exp(j\,k\,Sx)\,dx$$

$$= RI_o \cos\theta \int_{-L/2}^{L/2} \exp(j\,k\,Sx)\,dx$$

$$= RI_o \cos\theta . \frac{2\sin\dfrac{kSL}{2}}{kS} \quad \ldots \quad \ldots \quad (17)$$

This, as is to be expected, is exactly equivalent to the equation giving the frequency spectrum of a unit pulse, apart from the term in $\cos\theta$. Since $k = \dfrac{2\pi}{\lambda}$ the equation may be written in the form

$$P(S) = RI_o \cos\theta . \frac{2\sin\dfrac{SL\pi}{\lambda}}{\dfrac{2\pi S}{\lambda}} \quad \ldots \quad \ldots \quad (18)$$

and Fig. 1 gives the polar diagram of the aerial provided that

for $s(\omega)$ we read $P(S)$
for ω we read $S = \sin\theta$
for ω_0 we read λ/L

and provided further that we multiply each point on the diagram by the appropriate value of $\cos\theta$.

There is, in polar diagram case, a physical limitation which did not occur in the frequency spectrum case and that is that we shall not be interested in the values of S greater than unity, or less than -1. The values of $P(S)$ for $S > 1$ or $S < -1$, correspond physically to storage fields around the aerial, the energy in which returns to the aerial in alternate quarter-cycles, and this energy does not, therefore, contribute to the radiation field.

We notice from the diagram that the direction of maximum radiation is normal to the length of the aerial, and that the width of the main lobe depends upon the ratio of λ/L.

The modulation theorem, as applied to the aerial case may be stated thus:—

"If $P(S)$ is the polar diagram of an aerial with aperture distribution $f(x)$, then the polar diagram of the aerial with the modulated aperture distribution $f(x)\exp(-jkS_c x)$ is $P(S - S_c)$."

In words, the effect of modulating the rectangular aperture distribution with the function $\exp(-jkS_c x)$ is merely to shift the spectrum of the rectangular distribution along the S ($= \sin\theta$) axis and centre it on $S = S_c$; i.e., the polar diagram is slewed.

Modulated aperture distributions may be obtained by placing in front of the aerial an absorbing screen of the correct characteristics, but they do, in fact, arise much more naturally when we consider ordinary aerials along which the current distribution is not uniform. For instance, we know that if we feed an open-ended dipole with oscillatory current, reflection of the wave at the end of the wire gives rise to a standing-wave current distribution which is sinusoidal in form and of zero amplitude at the two ends of the aerial. This particular current distribution can be expressed as the difference of two functions of the form $\exp(jkS_c x)$ and $\exp(-jkS_c x)$ representing waves travelling in the positive and negative directions along the wire.

Suppose, for example, that we wished to calculate the polar diagram of the aerial shown in Fig. 11 where there are four current loops along the aerial and the current distribution is sinusoidal and therefore anti-symmetrical about $x = 0$.

The current distribution is of the form $\sin \frac{4\pi x}{L}$ and this may be expressed in the form

$$\sin \frac{4\pi x}{L} = \frac{1}{2j}\left\{\exp\ (jkS_c x) - \exp\,(-jkS_c x)\right\}$$

whence we find $S_c = \frac{2\lambda}{L}$ since $k = \frac{2\pi}{\lambda}$

Furthermore, the aerial is 2λ long, so that $\frac{\lambda}{L} = \frac{1}{2}$

To calculate the polar diagram we need to take half the difference of the polar diagrams of two unit rectangular aperture distributions centred on $S = \pm 1$ (for $S_c = \frac{2\lambda}{L} = 1$), and multiply the result by $\cos\theta$, to take into account the polar diagrams of the elementary doublets. We remember, too, that we need not calculate the polar diagram for values of S outside the range ± 1. The patterns to be added and their relative positions are shown in Fig. 12 in which the dotted line represents the resultant, and the polar diagram obtained by multiplying the resultant by the appropriate values of $\cos\theta$ is shown in Fig. 13.

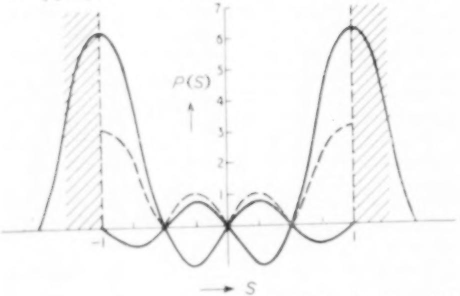


Fig. 12. *Component patterns of the polar diagram of the aerial of Fig.* 11.

We notice in Fig. 13 that the ratio of the major to the minor lobes is not particularly large. In general, of course, the directivity of an aerial may be improved by increasing its length. But suppose we are limited, by physical considerations, to an aerial which is two wavelengths long. Is there any simple way of decreasing the minor lobes? Now the subsidiary lobe of Fig. 13 was calculated by taking half the difference of the overlapping spectra of Fig. 12 in the range $S = 0$ to $S = 0{\cdot}5$. In this range of values of S, with the relative positions of the component spectra fixed, as shown, by the length of the aerial, the components of the two spectra are opposite in sign, and therefore, their difference is appreciable. Had we been able to form the resultant spectrum by taking half the sum of the two component spectra, then in the range $S = 0$ to $S = 0{\cdot}5$ the value would have been considerably decreased, while in the range $S = 0{\cdot}5$ to $S = 1{\cdot}0$ the value would have been much affected since the components of the overlapping spectra in this range are greatly different.

Thus, by taking half the sum, instead of half the difference, of the component spectra, we shall be able to improve the directivity of the aerial by eliminating almost completely the subsidiary lobes. In order to take half the sum of the component spectra, instead of half their difference, we shall need to have a current distribution with a current loop at the mid-point of the aerial instead of a current node. This can be arranged by short-circuiting the ends of the aerial instead of leaving them open as in Fig. 11. An effect similar to short-circuiting could be achieved by connecting to each end of the aerial a piece of wire a half-wavelength long, the connection being made at the centre of the stub, and the stub itself being set perpendicular to the length of the aerial. The current in the loops which now appear at each end of the aerial itself, divides and flows in opposite directions down the two halves of each stub, so that the actual radiation from the stubs themselves to distant points is small. In this way the aerial is effectively short-circuited at the ends, giving the desired decrease in the size of the minor lobes, while the radiation from the shorting stubs is reduced to a small value.

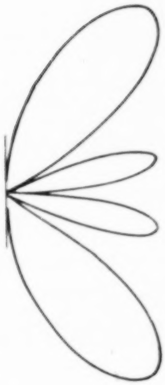


Fig. 13. *Final polar diagram of the aerial of Fig.* 11.

The case which we have just considered in Figs. 11, 12 and 13 is a simple one since the reflection coefficient to the open-ended wire is unity (negative) giving a distribution of current across the aperture which is sinusoidal and anti-symmetrical about $x = 0$. If the reflection coefficient is not unity, one of the component spectra must be multiplied by the appropriate constant before the addition is performed, and if any phase lag is introduced at reflection this must also be taken into account.

This method of calculation of the resultant is especially useful when the reflection coefficient is unity but the aerial is not an integral number of half wavelengths long. In such cases, the aperture distribution of current is analysed in exactly the same way as we have already employed in finding the frequency spectrum of a section of a cosine wave of arbitrary length and initial phase. A formula of the same form as equation (12) is easily developed and is just as useful. It may be written,

$$P'(S) = RI_0 \cos\theta \left[\left\{ \frac{\sin (S+S_c)\dfrac{\pi L}{\lambda}}{(S+S_c)} + \frac{\sin (S-S_c)\dfrac{\pi L}{\lambda}}{(S-S_c)} \right\}^2 - \frac{4 \sin (S+S_c)\dfrac{\pi L}{\lambda}}{(S+S_c)} \cdot \frac{\sin (S-S_c)\dfrac{\pi L}{\lambda}}{(S-S_c)} \cdot \sin^2\phi \right]^{\frac{1}{2}} \qquad (19)$$

where ϕ is the phase angle at the centre of the aerial ($x = 0$) of the aperture distribution regarded as a cosine wave.

If the reflection coefficient is not unity, the appropriate value must be placed in front of the term in $(S+S_c)$ in equation (19), and ϕ must also be modified if necessary.

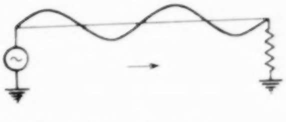


Fig. 14 (above). *Two wavelength travelling-wave aerial.*

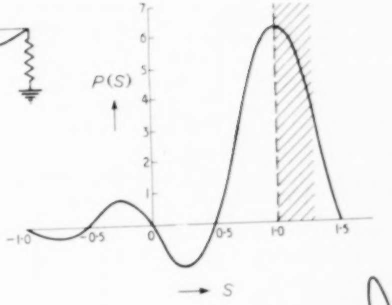


Fig. 15 (right). *Angular spectrum for the aerial of Fig. 14.*

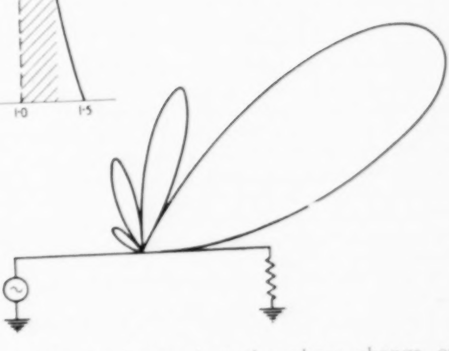
5. Travelling-Wave Aerials

The method may be used for the calculation of the polar diagrams of travelling-wave aerials; i.e., aerials which are correctly terminated so that there is no reflected wave. For this case we may notice that the term in $(S+S_c)$ in equation (19) which corresponds to the reflected wave disappears, so that we are left only with the polar diagram of the rectangular distribution centred on S_c. We can see this equally well by remembering that exp $(-jkS_cx)$ represents a wave travelling in the positive x-direction.

For instance, suppose that we wish to find the polar diagram of a travelling-wave aerial 2λ long as shown in Fig. 14. At the particular instant shown, the distribution of current along the aerial may be written sin $4\pi x/L$ which may be expressed in the form exp $(-jkS_cx)$, where $S_c = 2\lambda/L = 1$ since $\lambda/L = \frac{1}{2}$. The polar diagram of the travelling-wave aerial is thus given by the angular spectrum of the unit rectangular aperture distribution centred on $S = 1$, as shown in Fig. 15, multiplied by cos θ. The resulting polar diagram is shown in Fig. 16.

By this method of approach to the calculation of aerial polar diagrams, the connection between the polar diagram of a resonant aerial of given length and that of a travelling-wave aerial of the same length becomes apparent. In particular, we notice that if the aerials are a few wavelengths long the polar diagram of the resonant aerial is practically the same as that of the travelling-wave aerial together with its reflection in the $\theta = 0$ line. A few minor modifications are introduced in the subsidiary lobes close to the $\theta = 0$ axis, but the main lobes are practically identical. If, under the same conditions (i.e., a long aerial, $\lambda/L \ll 1$) the reflection coefficient for the resonant case is not unity, the portion of the polar diagram reflected in the $\theta = 0$ axis must be multiplied by the appropriate factor, but the phase change on reflection is not important and may be neglected except when dealing with minor lobes near the $\theta = 0$ axis.

Fig. 16 (below). *Polar diagram of the aerial of Fig. 14.*



6. Conclusion

Throughout this work, attention has been confined to rectangular envelopes; e.g., rectangular pulses and rectangular aperture distributions.

The same methods and ideas may, however, be applied to the other enveloping functions. It is necessary, for the easy application of the graphical method and the trigonometrical formula that the frequency spectrum of the enveloping function should be expressible as the sum of a series of sine waves, or of cosine waves, which are all in phase at time $t = 0$ or at point $x = 0$. Just one example of such a function is the symmetrical triangular function, which might be used as an approximation in dealing with an aerial with loss along its length.

As was pointed out in the introduction, this investigation was begun because of the errors found when the effects of negative frequencies were too lightly dismissed but it has led on to general points in Fourier analysis which, it is hoped, may prove interesting and instructive. The idea of introducing an auxiliary function has parallels in other branches of mathematics as, for example, in the integration of the function $\log_e x$, where the process is made much simpler by introducing the auxiliary function unity, and integrating the product $1 \times \log_e x$ by parts. The introduction of the rectangular pulse leads to similar simplification in many cases of Fourier analysis as is shown here, and there is no doubt that many other examples of its use may be found.

CATHODE-COUPLED PULSE GENERATOR

With Square-Wave Control

By F. A. Benson, M.Eng., A.M.I.E.E., M.I.R.E., and G. V. G. Lusher, B.Eng.

(The University of Sheffield)

SUMMARY.—A two-valve pulse generator with a fixed repetition rate is described. The positive pulses which it produces have a time of rise of slightly less than 0·5 microsecond, an amplitude of approximately 30 V, and a width of about 1 microsecond.

Introduction

THE circuits most commonly used for the production of pulses at a fixed repetition rate use the principle of 'clipping' a sine wave and passing the 'clipped' wave through an RC differentiating circuit. This method has the disadvantage that the time of rise is dependent on how near the 'clipped' wave approaches a true square wave. If a time of rise of less than 1 microsecond is required it is necessary that the clipper output be amplified again and re-clipped. The pulses obtained in this way have a gradual exponential fall. An alternative method of producing negative pulses has been described recently by Benson and Pearson.[1] In this case the pulses have a time of rise of 0·8 microsecond, have a gradual decay and are about 60 V in amplitude. It does not appear possible to reduce this time of rise with the circuit suggested.

The present article gives a further method of producing positive pulses using a highly-damped oscillatory circuit in a cathode-coupled flip-flop arrangement. The pulses obtained have a time of rise of a little less than 0·5 microsecond; they are approximately 30 V in amplitude and have a width of about 1 microsecond.

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The use of a highly-damped oscillatory circuit for the production of a short pulse from a sudden voltage change may possibly be well known and the method is illustrated in Fig. 1. Suppose a positive-going voltage change is applied through a capacitor C to a parallel circuit L, C_s and R which has a diode V across it. C_s represents the sum of the capacitance of the diode and the stray capacitance of coil L. This voltage change 'kicks' the oscillatory circuit LC_s, into oscillation. If the damping resistor R and the diode V are both disconnected, the waveform shown in Fig. 2(a) is obtained. With the damping resistor R connected the oscillation is damped and the waveform of Fig.

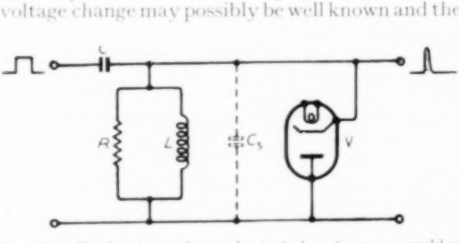


Fig. 1. *Production of a short pulse from a sudden positive-going voltage change, with the aid of a highly-damped oscillatory circuit.*

2(b) results. If, in addition, the diode is connected, the diode conducts on the negative part of the cycle and, hence, the negative portion of the waveform is removed as illustrated in Fig. 2(c)

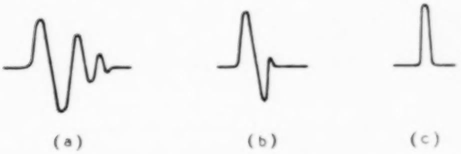


Fig. 2. *Waveforms in the circuit of Fig. 1; (a) without R or V, (b) with R and (c) with R and V.*

Fig. 3. *Production of a short pulse from a sudden negative-going voltage change, with the aid of a highly-damped oscillatory circuit.*

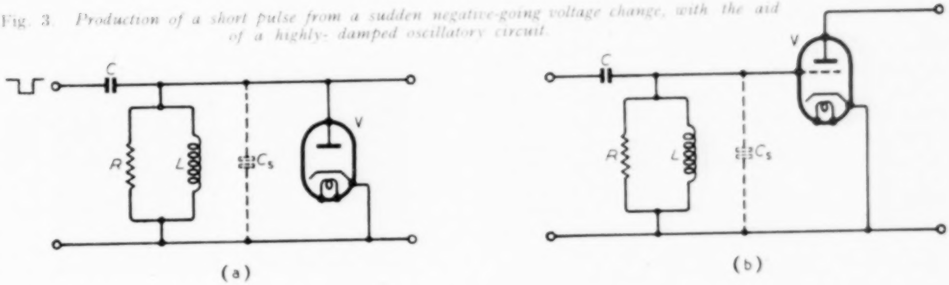


Fig. 4 (right). *Circuit arrangement of cathode-coupled flip-flop pulse generator.*

The steepness of the rise depends on the steepness of the input-voltage change, and the steepness of the fall and the duration of the pulse depend on the frequency of the LC_s circuit.

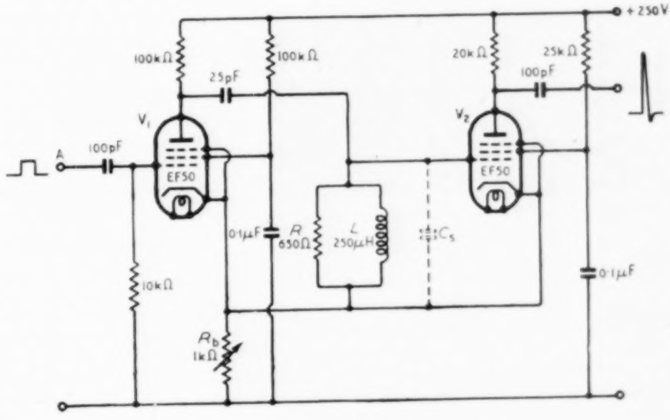
If a negative-going voltage change is to be used instead of a positive one as discussed above, the diode must be connected as shown in Fig. 3(a) or, alternatively, a triode may be employed giving some additional amplification [Fig. 3(b)]. In the latter case, C_s is composed of the grid-cathode capacitance of the triode and the stray capacitance of the coil L and grid-current effects are the same as those produced by diode current in the previous circuits.

By using such a highly-damped oscillatory circuit in a cathode-coupled flip-flop arrangement a much sharper pulse may be obtained as described below.

Flip-Flop Circuit

The circuit diagram is shown in Fig. 4. The valves used are pentodes type EF50 connected as a conventional cathode-coupled flip-flop, except that the normal grid resistor of V_2 is replaced by a highly-damped oscillatory circuit L, C_s, R. R is a variable damping resistor and C_s is the sum of the grid-cathode capacitance of valve V_2 and the self-capacitance of coil L.

A positive voltage change is required at the input. The authors obtained this from a multivibrator giving a voltage of about 50 V and operating at a frequency of 6·5 kc/s.



Explanation of the Circuit Action

Consider the positive voltage change from the multivibrator to be applied at point A (Fig. 3). This causes valve V_1 to pass from the cut-off stage and, hence, its anode voltage falls rapidly. This is the first stage of the flip-flop action which drives the grid of V_2 well beyond cut-off. At the same time the tuned grid circuit is set into oscillation. The grid of V_2 returns to cathode voltage in a time determined by the natural frequency of the *L*, *C*, and *R* circuit. As the grid voltage of V_2 reaches the point at which the anode current of that valve begins to flow again, the second phase of the flip-flop action is initiated and the original circuit conditions are restored in readiness for the next positive voltage change.

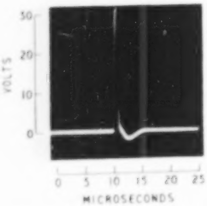


Fig. 5. *Typical voltage pulse.*

During the second phase of the flip-flop action the grid of V_2 is driven positive, and the tuned circuit is highly damped. The action is the same as for the diode circuit illustrated in Fig. 1(c). The oscillatory current is, therefore, damped out after the grid has completed one negative excursion.

Due to the flip-flop action of the circuit, the grid of V_2 is driven slightly positive with respect to the cathode and, therefore, the final edge of the pulse from the anode of V_2 has a small negative excursion. This is clearly seen in Fig. 5 which shows a typical output pulse. The negative portion of the pulse can be reduced to a minimum by viewing the waveform on an oscilloscope and adjusting the common cathode resistor.

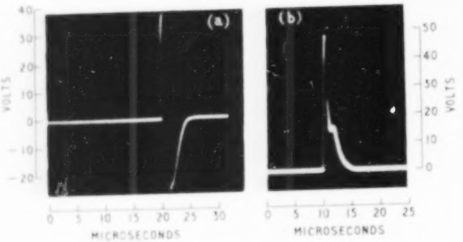


Fig. 6. *Typical voltage pulses showing effects of varying cathode-bias resistance R_b.*

Effects of Varying Component Values

The component values shown in Fig. 4 are designed for an h.t. supply of 250 V and, apart from the cathode-bias resistor R_b and the coil, are not critical. The effects of varying R_b are shown in Fig. 6. The large negative part of the waveform in Fig. 6 (a) is produced when R_b is greater than 650 ohms, but when R_b is less than this value the waveform in Fig. 6 (b) is obtained.

The effects produced by employing different coil inductances are worth noting. With a 15-μH coil, and circuit values shown in Fig. 4, a waveform having a very steep rise (0·4 microsecond), but a jagged rear edge is obtained. Fig. 7 (a) shows such a waveform. The amplitude of the pulse is reduced to 16 V. With a 400-μH coil the waveform of Fig. 7 (b) is obtained.

In this case the amplitude increases to about 50 V but the time of rise also rises to approximately 1 microsecond. The value of coil inductance given in Fig. 4 appears to be the most satisfactory for obtaining a short pulse of good amplitude.

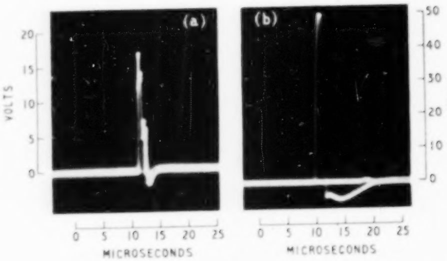


Fig. 7. *Typical voltage pulses showing effects of varying coil inductance L.*

Conclusion

The simple pulse generator described, which can be built from a few normal laboratory components, gives short positive pulses of good amplitude which cannot be conveniently produced by 'clipping' and differentiating circuits.

Acknowledgement

The work recorded in this paper has been carried out in the Department of Electrical Engineering at the University of Sheffield. The authors wish to thank Mr. O. I. Butler, M.Sc., M.I.E.E., for the facilities afforded in the laboratories of this Department.

REFERENCE

[1] Benson, F. A., and Pearson, R. M., *Wireless Engineer*, December 1950, Vol. 27, pp. 285-288.

PULSE RESPONSE OF A.M. RECEIVER

By R. Kitai. M.Sc.(Eng.)(Rand)

SUMMARY.—It is shown that, when a square pulse of voltage of short duration is induced in the aerial of an amplitude-modulated receiver, the magnitude of the detector output voltage is dependent on the *duration* of the pulse as well as the pulse amplitude. This means that in many cases the simple superheterodyne receiver can be used to determine, with good accuracy, the duration of periodic short pulses. Theory also shows that for impulsive interference, the amplitude of the response of the receiver can depend to a large measure on the frequency to which the receiver is tuned.

IN this analysis the transient response of a small portion of the simple amplitude-modulated receiver alone will be considered, it being assumed that the receiver is of the superheterodyne type without a radio-frequency amplifier stage, and that the aerial coupling circuit is of the common inductive type used in conjunction with an open-ended aerial of the order of 4 metres effective height. As will be seen, the mathematical analysis of the receiver beyond the aerial coupling circuit need not concern us here, but it is important to point out that in the simple superheterodyne receiver, the shape of the transient voltage fed to the power-amplifier valve is dependent to a large measure on the characteristics of the i.f. transformer of the narrowest bandwidth. Usually, then, the final shape of the transient voltage fed to the power amplifier bears little relation to the shape of the transient voltage induced in the aerial; furthermore the shape of the transient voltage at the input to the power-amplifier valve is hardly affected by variation of the frequency to which the receiver is tuned.

With regard to the transient *amplitudes*, however, the above does not apply, and although the i.f. circuit 'shapes' the transient output, the amplitude of this output is proportional to the amplitude of the transient voltage induced in the aerial, provided the amplitude characteristics of the receiver are sensibly linear within the range considered.

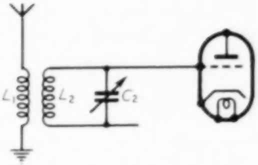


Fig. 1. *Aerial tuning circuit of receiver.*

In analysing the transient response of the aerial coupling circuit a Heaviside step voltage $E.\mathbf{1}$ is assumed to be induced in the aerial, the circuit being shown in Fig. 1. An aerial with an effective height of 4 metres or thereabouts usually has a capacitance to earth in the range 100-300 pF and a primary inductance L_1 of one or two millihenrys is chosen for the broadcast band to ensure a primary resonant frequency below 550 kc/s. A coupling of 10% furnishes enough mutual inductance to give adequate signal transfer to the secondary circuit, and a resistance of the order of 25 ohms is added to the primary resistance of the coil to account for the aerial resistance.

Analysis of the Circuit

The circuit to be analysed is given in Fig. 2 where the aerial is replaced by the capacitance C_1 and a portion of the resistance R_1.

Using the Heaviside operator the differential equations for primary and secondary circuits are:—

$$\left(pL_1 + R_1 + \frac{1}{pC_1}\right)i_1 + pMi_2 = E.\mathbf{1}$$

$$\left(pL_2 + R_2 + \frac{1}{pC_2}\right)i_2 + pMi_1 = 0$$

and

$$v_2 = \frac{1}{pC_2}\,i_2$$

On eliminating i_1 and substituting

$$\alpha_1 = \frac{R_1}{2L_1},\ \alpha_2 = \frac{R_2}{2L_2},\ \omega_1{}^2 = \frac{1}{L_1C_1},\ \omega_2{}^2 = \frac{1}{L_2C_2}$$

$$\text{and } k = \frac{M}{\sqrt{L_1L_2}}$$

the expression for the secondary terminal voltage becomes

$$-\frac{L_1L_2C_2}{p^2M}\Big\{(1-k^2)\,p^4 + 2\,(\alpha_1+\alpha_2)\,p^3 + (\omega_1{}^2 + 4\alpha_1\alpha_2 + \omega_2{}^2)p^2 + 2(\alpha_1\omega_2{}^2 + \alpha_2\omega_1{}^2)\,p + \omega_1{}^2\omega_2{}^2\Big\}v_2 = E.\mathbf{1} \qquad \text{.. .. ..} \quad (1)$$

When the circuits are highly oscillatory, the roots of p in this fourth-degree differential equation may be obtained by a method due to Guillemin.[1] The solution is an approximation obtained by assuming initially that the circuits have no damping. The odd powers of p then

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disappear in (1) equation which now becomes a biquadratic and yields four solutions in the form of two purely imaginary conjugate pairs. Thus, if equation (1) is written

$$-\frac{L_1L_2C_2}{Mp^2}\left\{ap^4+bp^3+cp^2+dp+f\right\}v_2=E.\mathbf{1}$$

then if R_1 and $R_2 \to 0$, b and $d \to 0$

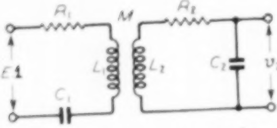

Fig. 2. *Equivalent circuit of Fig.* 1.

Equating the bracketed term to zero and writing p_1 as the approximate roots,

$$\alpha p_1^{\,4}+cp_1^{\,2}+f=0$$

whence $$p_1^{\,2}=\frac{-c\pm\sqrt{c^2-4af}}{2a}$$

Now for k of the order 0·1 or less, $1-k^2$ is very nearly unity so that the determinant $\sqrt{c^2-4af}$ is very nearly $\omega_1^{\,2}-\omega_2^{\,2}$ giving

$$p_1=\pm j\omega_1 \text{ or } p_1=\pm j\omega_2.$$

These approximate roots are now used in Newton's approximation method to obtain a more exact solution. Thus if δ is the first correction term, then

$$\delta=-\frac{ap_1^{\,4}+bp_1^{\,3}+cp_1^{\,2}+dp_1+f}{4ap_1^{\,3}+3bp_1^{\,2}+2cp_1+d}$$

$$\approx-\frac{bp_1^{\,3}+dp_1}{4ap_1^{\,3}+2cp_1}=-\frac{bp_1^{\,2}+d}{4ap_1^{\,2}+2c}$$

Now when $p_1=\pm j\omega_1$

$$\delta=\delta_1=-\frac{-2(\alpha_1+\alpha_2)\omega_1^{\,2}+2(\alpha_1\omega_2^{\,2}+\alpha_2\omega_1^{\,2})}{-4\omega_1^{\,2}+2(\omega_1^{\,2}+\omega_2^{\,2})}=-\alpha_1$$

and similarly, when

$$p_1=\pm j\omega_2$$
$$\delta=\delta_2=-\alpha_2$$

so that the roots of p are very nearly

$$p=-\alpha_1\pm j\omega_1,\quad p=-\alpha_2\pm j\omega_2 \quad .. \qquad (2)$$

Since the circuits are highly oscillatory it is not usually necessary to obtain more accurate roots, and the final expression for v_2 can be found by substituting these roots in Heaviside's Expansion theorem. In using this theorem, equation (1) is expressed in the form

$$v_2=\frac{1}{Z(p)}E.\mathbf{1}$$

whence the solution for v_2 is given by

$$v_2=\left\{\frac{1}{Z(0)}+\sum\frac{\epsilon^{pt}}{pZ'(p)}\right\}E \quad .. \quad .. \qquad (3)$$

In substituting the roots of p found above, an approximate solution is obtained by neglecting the small reals in the expression for $pZ'(p)$. Thus if decrement terms are neglected,

$$Z(p)=-\frac{L_1L_2C_2}{M}\left(ap^2+c+\frac{f}{p^2}\right)$$
$$=-\frac{L_1}{\omega_2^{\,2}M}\left(ap^2+c+\frac{f}{p^2}\right)$$

so that

$$pZ'(p)\approx-\frac{2L_1}{M}\left(\frac{p^2}{\omega_2^{\,2}}-\frac{\omega_1^{\,2}}{p^2}\right) \quad .. \qquad (4)$$

where $k^2 \leqslant 1$.

The values of p given by equation (2) are now regarded as pure imaginaries when substituted into equation (4). These approximations give results sufficiently accurate for most practical cases where ω_1 is neither equal to nor very nearly equal to ω_2, so that equations (2), (3) and (4) give

$$v_2\approx\frac{ME}{2L_1\left[1-\left(\frac{\omega_1}{\omega_2}\right)^2\right]}$$

$$\left\{\epsilon^{(-\alpha_1+j\omega_1)t}+\epsilon^{(-\alpha_1-j\omega_1)t}-\epsilon^{(-\alpha_2+j\omega_2)t}-\epsilon^{(-\alpha_2-j\omega_2)t}\right\}$$

or, writing $\gamma=\left(\frac{\omega_1}{\omega_2}\right)^2$

$$v_2\approx\frac{ME}{L_1(1-\gamma)}\left\{\epsilon^{-\alpha_1 t}\cos\omega_1 t-\epsilon^{-\alpha_2 t}\cos\omega_2 t\right\} \quad (5)$$

The voltage on the signal grid of the convertor valve is thus seen to consist of the two exponentially-decaying oscillations at the primary and secondary resonant frequencies. Referring back to Fig. 1, it is seen that ω_1 is fixed by the inductance L_1 of the coupling coil and by the aerial capacitance C_1, while ω_2 varies with variation of the tuning capacitance C_2. Now in the simple superheterodyne receiver a comparative uniformity of steady-state response to all frequencies within any particular band is achieved by ensuring that any oscillation at the aerial-circuit resonant frequency is well outside the band of frequencies to which the receiver can be tuned. Usually ω_1 is made to lie below the lowest value of ω_2 for each band so that under these circumstances γ is always less than unity and is usually small compared with unity.

With regard to the convertor local oscillator and the first i.f. transformer, these are therefore tuned to accept the oscillation at ω_2 radians/sec and to reject the oscillation at ω_1 radians/sec. In examining the response of the receiver from the convertor valve onwards, it is now only necessary to consider the second term in the bracket of equation (5).

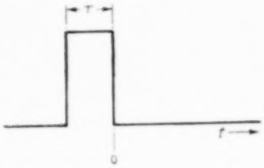

Fig. 3. *Square wave of duration τ.*

Let us examine the response of the r.f. circuit of a receiver to the square pulse shown in Fig. 3, the duration of the pulse being τ seconds, the time origin being taken at the downward end of the pulse.

From equation (5) the response of the circuit t seconds after zero time is

$$v_2 = -\frac{ME}{L_1(1-\gamma)}\left\{\epsilon^{-\alpha_2(t+\tau)}\cos\omega_2(t+\tau) - \epsilon^{-\alpha_2 t}\cos\omega_2 t\right\}$$

or

$$v_2 = \frac{ME}{L_1(1-\gamma)}\,\epsilon^{-\alpha_2 t}\left\{\cos\omega_2 t - \epsilon^{-\alpha_2\tau}\cos\omega_2(t+\tau)\right\} \quad \ldots \quad \ldots \quad (6)$$

This equation shows that when

$$\omega_2\tau = \pi, 3\pi, 5\pi \ldots.$$

then

$$v_2 = \frac{ME}{L_1(1-\gamma)}\left[1+\epsilon^{-\alpha_2\tau}\right]\epsilon^{-\alpha_2 t}\cos\omega_2 t\,,$$

the responses due to the steps in the pulse being cumulative. Similarly, when

$$\omega_2\tau = 0, \pi, 2\pi \ldots.$$

then

$$v_2 = \frac{ME}{L_1(1-\gamma)}\left[1-\epsilon^{-\alpha_2\tau}\right]\epsilon^{-\alpha_2 t}\cos\omega_2 t\,,$$

the responses due to each step being differential.

An illustration showing the variation in amplitude of v_2 with pulse duration τ for a particular case is shown in Fig. 4. The receiver is assumed to be tuned to a fixed frequency of 1 Mc/s, and α_2 has been chosen as 7×10^4. (This value is representative of most receivers where Q_2 is of the order 50.) Thus a pulse of duration one microsecond corresponds to a period of the oscillation of the secondary tuned circuit. This considerable variation of v_2 with pulse duration can easily be demonstrated on a cathode-ray oscilloscope connected to the detector output of a receiver, the receiver aerial being situated near to, or connected to, a generator having an output consisting of a periodic square pulse of variable duration.

Returning to equation (6), it is also evident that for a fixed pulse duration, the amplitude of v_2 varies with receiver tuning. Thus, for impulsive interference, the final amplitude of the response of the receiver can depend to a large measure on the frequency to which the receiver is tuned. This matter becomes clear from a consideration of the next section dealing with the measurement of the duration of square pulses. It will be seen that, for pulses of rather short duration, the longer the duration of the pulse, the greater the fluctuation of the receiver response with tuned frequency.

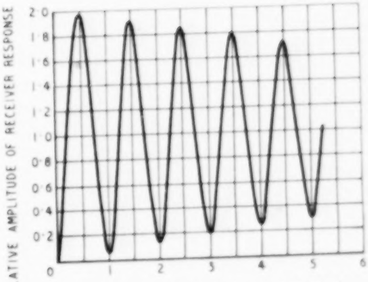

Fig. 4. *Variation of amplitude with pulse duration.*

Measurement of Pulse Duration by means of a Receiver

A receiver can often be used to measure the duration of square repetitive pulses with good accuracy and to adjust a short repetitive pulse to have a required duration. As an illustration, let us take the case of a square-pulse generator having any constant pulse-recurrence frequency, and let it be desired to adjust the duration of the pulse to one microsecond. If the generator is connected to the input of the receiver tuned to 1 Mc/s, then the periodic time of the receiver r.f. circuit oscillation is one-microsecond. The rising and falling sides of a one microsecond square pulse would then act differentially so that the voltage at the output of the detector would be a minimum. If the pulse is of one microsecond duration and the receiver tuning altered, the next minimum response will occur when the r.f. circuit has two cycles of oscillation during a period of one microsecond; i.e., the second minimum response will occur at a tuned frequency of 2 Mc/s, the third minimum at 3 Mc/s, and so on, *irrespective of the pulse recurrence frequency.* Fig. 5 shows the results obtained with a receiver in which the a.g.c. had been switched out of circuit. The dotted curve is the envelope of the line-frequency spectrum given by the Fourier equation for a repetitive square pulse of short duration.[2]

The lower limit of pulse-recurrence frequency is set by the time available for making the measurements. For high p.r.fs the individual line frequencies of the Fourier analysis will be distinguishable as the receiver tuning is varied; the wave-shape at the detector output is also likely to vary at high p.r.f. due to the overlapping of comparatively slowly-decaying transients in the i.f. transformers. These, however, will not affect the accuracy of the experiment.

Next, a practical example is chosen to illustrate how the duration of a repetitive square pulse can be determined. In this experiment the receiver tuning was varied over the broadcast band with a.g.c. in circuit, and response minima in the detector output were observed at frequencies of 0·78, 1·04, 1·31, 1·56 and 1·82 Mc/s for a square pulse of unknown duration and p.r.f. = 500 c/s. The receiver was of the communication type with a well-graduated scale, and was in good alignment. Minima alone were noted, since the response curve has rapid rates of change in the vicinity of these minima. As a check on the alignment of the receiver it is observed that the frequency separation between successive minima are respectively 0·26, 0·27, 0·25 and 0·26 Mc/s.

Greatest accuracy is obtained if the lowest and highest of these frequencies are used in the calculations. Thus, let the duration of the pulse be τ microseconds and at 0·78 Mc/s let there be n cycles of transient oscillation of the r.f. circuit for the duration τ. Then at a frequency of 1·82 Mc/s there will be $(n + 4)$ cycles of oscillation of the r.f. circuit.

$$\text{Hence } \tau = \frac{n}{0{\cdot}78} = \frac{n+4}{1{\cdot}82} \text{ microseconds}$$

whence $n = 3{\cdot}00$ cycles
and $\tau = 3{\cdot}85$ microseconds.

The fact that n is very nearly a whole number shows that τ has been measured with good accuracy. If the receiver calibration is doubtful, the frequencies at the receiver-dial settings for response minima can be obtained more accurately with the aid of a signal generator.

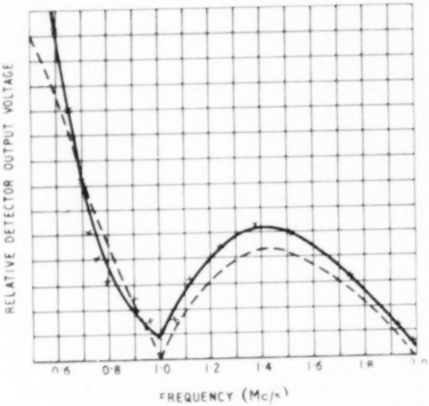

Fig. 5. *Variation of receiver output with tuning.*

The examples show that for a one-microsecond pulse, successive response minima exist at frequencies of 1 and 2 M/cs while for a pulse of duration 3·85 microseconds, four response minima exist between these two frequencies. It is thus clear that the longer the duration of the pulse, the greater the fluctuation of the receiver response with variation of tuned frequency.

REFERENCES

[1] Guillemin, E. A. "Communication Networks." 1931. Vol. 1, p. 206, John Wiley & Sons, New York.
[2] Terman, F. E. "Radio Engineer's Handbook." 1943, pp. 21-22 McGraw-Hill Book Co., New York.

I.E.E. MEETINGS

8th January. "Two Electronic Resistance or Conductance Meters", by L. B. Turner, M.A., Sc.D., and "A Bridge for the Measurement of Dielectric Constants of Gases", by W. F. Lovering, M.Sc., and L. Wiltshire, M.Sc.

16th January. "Comparison of Ionospheric Radio Transmission Forecasts with Practical Results", by A. F. Wilkins, O.B.E., M.Sc., and C. M. Minnis, M.Sc.

23rd January. Discussion on "Essentials of a First Course in Electricity and Magnetism", opened by H. Kayser, B.Sc.

28th January. Discussion on "Should Further Television Development be Concentrated on Colour to the Exclusion of Black and White?", opened by L. C. Jesty.

All meetings are to be held at the Institution, Savoy Place, London, W.C.2, at 5.30 p.m. with the exception of the meeting on 23rd January, which commences at 6 p.m.

BRIT.I.R.E. MEETING

9th January. "Crystal Triodes", by E. G. James, Ph.D., and G. M. Wells, B.A. To be held at the London School of Hygiene and Tropical Medicine, Keppel St., Gower St., London, W.C.1, at 6.30 p.m.

WIDEBAND PRE-AMPLIFIER

For Atmospheric-Noise Measurement

By F. Horner, M.Sc., A.M.I.E.E.

(Communication from the National Physical Laboratory)

SUMMARY—A wideband pre-amplifier has been designed to improve the sensitivity of existing equipment used for measuring atmospheric-noise levels. The unit is associated with the aerial and operates unattended at a point remote from the rest of the equipment. The method of measurement is the aural comparison of a locally-generated morse signal with the noise picked up by the aerial, and means are provided for injecting the comparison signal into the first stage of the pre-amplifier.

The most difficult design problem is the avoidance of intermodulation between received signals, due to non-linear amplification, since the generation of spurious signals reduces the number of clear channels available for noise measurements and may also lead to erroneous readings. The effect has been reduced by restricting the response to the required frequency band (2·5–20 Mc/s), by careful design of the final cathode-follower stage, and by making the voltage gain no greater than is required to achieve the desired sensitivity.

The sensitivity is such that with a receiver of 10 kc/s bandwidth, and in the absence of atmospherics, a c.w. signal with a field strength of about 0·05 μV/m is intelligible in the presence of set noise only.

1. Introduction

IN a recent paper[1] a simple subjective method for measuring atmospheric noise has been described. The method is based on the aural comparison of a locally-generated morse signal with the noise, and has been used for routine observations in the band 2·5–20 Mc/s.

One consideration which influenced the design of the equipment was that noise levels less than 1 μV/m were not considered to be of practical interest, and some sacrifice of sensitivity was made to obtain greater simplicity of equipment and its operation.

When measurements were started, it became evident that noise levels were much below 1 μV/m at many stations for a large part of the time, especially at the higher frequencies, and that even these low values were of practical importance. Consideration was therefore given to various possible methods of increasing sensitivity. The paper describes a wideband pre-amplifier which was developed for this purpose and deals particularly with the intermodulation problems which arise in such an amplifier.

2. The Original Equipment

The method of measurement is the aural comparison of the noise, picked up by the aerial, with a locally-generated morse signal of controllable amplitude. The basic circuit of the original equipment is shown in Fig. 1, which is reproduced from reference 1. Except at 10 Mc/s, where the aerial is resonant, sensitivity is low, due mainly to the 60-ohm resistor connected between the aerial and earth to match the cable. Other disadvantages of the arrangement shown are—

(a) since the aerial impedance is in general much larger than the impedance of the cable and terminating resistance, the voltage applied to the cable is critically dependent on the aerial impedance, which must, therefore, be accurately known and also independent of ground constants.

(b) Although the receiver input impedance is nominally 200 ohms it varies with frequency, and the voltage applied to the receiver from the signal generator depends on this impedance.

(c) When the aerial is resonant its impedance is about 40 ohms, and the cable is not correctly terminated at the aerial end. The transferred impedance at the receiver end of the cable is then not 60 ohms and an error is introduced in the estimated voltage applied to the receiver from the signal generator.

It must be emphasized that these defects in the circuit are unlikely to cause large errors in the measurements. Nevertheless, it is desirable to avoid them as far as possible in any modified circuit designed to improve the sensitivity.

3. Basic Requirements of the Modified Equipment

It was considered that the modified equipment should conform to the following basic requirements.

(a) As high sensitivity as possible and certainly high enough for noise fields of 0·1 μV/m to be measurable (see Section 4.1).

(b) Use of a remote aerial to avoid distortion of the polar diagram by the building housing the main equipment and to minimize interference from any man-made noise generated in the building.

(c) Simplicity of operation; in particular the avoidance of any tuning controls at the base of the aerial.

(d) Avoidance of critical dependence of the calibration on frequency.

(e) Effective bandwidth about 10 kc/s.

(f) Reliability.

In addition, it was desired to improve the performance with the minimum alteration to the existing equipment.

Various methods of achieving the desired sensitivity were considered, both theoretically and experimentally. Most of these involved tuning or loading the aerial and were ruled out because they introduced critical dependence of the calibration

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on frequency and remote switching of the circuits attached to the aerial. It was decided that the requirements could best be met by the use of a wideband pre-amplifier at the base of the aerial, covering the whole of the frequency range 2·5 to 20 Mc/s without tuning or switching. The major part of the development of such a pre-amplifier was carried out by H. A. Thomas, then on the staff of the National Physical Laboratory, in collaboration with staff of the National Bureau of Standards, Washington D.C., U.S.A., and was completed by the author at the N.P.L.

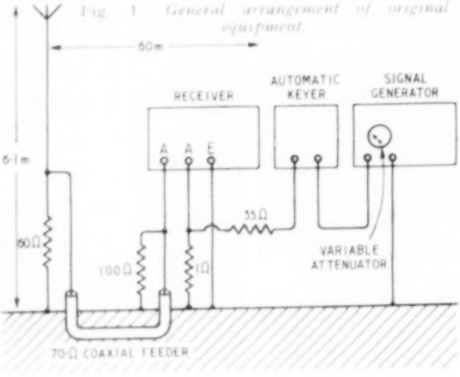


Fig. 1. General arrangement of original equipment.

4. Required Characteristics of the Pre-Amplifier

4.1. *Sensitivity*

It has been stated, somewhat loosely, that the equipment should permit measurement of noise fields of 0·1 μV/m, and it is desirable to consider more precisely what is meant by this statement. The sensitivity is limited by the inherent set noise, and to obtain a true measure of atmospheric noise by the comparison method described, it is necessary for the receiver output due to the noise field to be substantially greater than that due to the set noise; the observer cannot normally differentiate aurally between pure atmospheric noise and a mixture of atmospheric and set noise. The lower limit for reasonably accurate measurements occurs when the r.m.s. total-noise voltage is twice that of the internal set noise; the error in assuming that the noise is all picked up by the aerial is then quite small (15%).

The noise levels measured by the comparison method are expressed not as the field strength of the noise itself but as the field strength of a slow-speed morse signal (10 words per minute) necessary to provide 95% intelligibility in the presence of the noise, the bandwidth being 10 kc/s. Experiments have shown that the required field strength is about one-half the r.m.s. value of the noise field.

Thus when the lowest noise field which can be measured accurately is expressed in terms of the required c.w. field strength, its value is approximately equal to the equivalent r.m.s. field strength of the set noise. It is this quantity which is required to be less than 0·1 μV/m.

4.2. *Gain and Linearity*

To avoid tuning the pre-amplifier and to enable measurements to be made at any frequency in the band 2·5–20 Mc/s, the pre-amplifier is required to have uniform gain over this frequency band. To obtain high sensitivity the design should be such that the internal noise of the equipment is mainly derived from the first stage, and this should have low-noise characteristics. The pre-amplifier must have sufficient gain to ensure that the inherent noise of the main receiver is small compared with that arising in the pre-amplifier, and to ensure this a voltage gain of at least 2 from the input grid of the pre-amplifier to the input circuit of the main receiver is required.

Now a pre-amplifier of uniform gain over the range 2·5–20 Mc/s accepts a large number of signals, many of which have high field strengths, and it is difficult to avoid intermodulation due to non-linear amplification.

It is necessary to consider four possible effects of non-linearity.

(a) Various signals may beat together and generate a multitude of spurious signals which increase the difficulty of finding clear channels for noise measurements.

(b) Harmonics of individual signals may be produced, with the same effect as in (a).

(c) Strong signals may beat with components of the noise field and artificially increase the apparent noise level.

(d) Various components of the noise may beat together with the same effect as in (c).

Of these (a) is the most difficult to avoid because of the high intensity of the strongest signals compared with noise. Also the number of spurious signals tends to increase as the square of the number of genuine signals (considering all frequencies and not just the required pass-band), and so may easily become very large indeed. However, even when there is considerable intermodulation it is found that there are only a few strong spurious signals and that these do not cause serious interference with measurements since they can be recognized and avoided. A more serious effect is that in the background there are a very large number of small spurious signals, having approximately the same intensity as the noise and completely filling the spectrum. Moreover, these signals are not coherent and might easily be mistaken for genuine noise. It is this effect which must be avoided by reducing intermodulation to a minimum.

It is particularly important to avoid intermodulation involving components of the noise itself, mentioned in (c) and (d) above. This would artificially increase the noise level, and the measurement would be erroneous. However, the noise level is in general very low compared with the strong signals and if intermodulation between signals can be reduced to an acceptable level, intermodulation involving noise components is unlikely to be significant.

Ideally the equipment should accept two signals of high field strength (say 10 mV/m) and amplify them without introducing spurious signals greater than the inherent set noise level. This requirement is very stringent and almost certainly cannot be met in a practical circuit. Moreover, some intermodulation can be tolerated, but the amplification should be as linear as possible up to the point at which a narrow bandwidth is selected (the first stage of the main receiver). To maintain linear amplification the voltage gain should be no greater than is necessary to obtain high sensitivity. The gain should be as small as possible at all frequencies outside the required band 2·5–20 Mc/s.

4.3. *Comparison of the Artificial Signal with the Noise*

Means must be provided for injecting the locally-generated signal at some point in the receiver for comparison with the noise picked up by the aerial. To minimize the effects of possible variations in the equipment, the comparison should be made as near the aerial as possible, preferably in the first stage of the pre-amplifier.

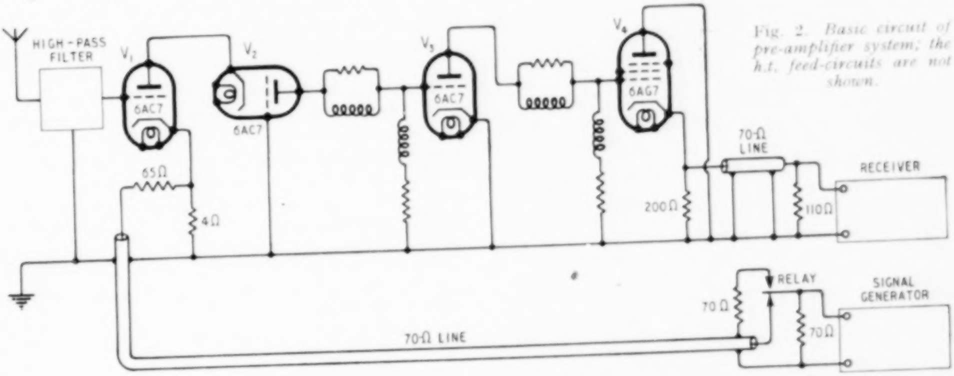


Fig. 2. *Basic circuit of pre-amplifier system; the h.t. feed-circuits are not shown.*

4.4. *Input and Output Impedance*

To obtain uniform performance over the frequency band without the use of matching circuits, the input impedance of the amplifier should be high compared with that of the aerial, so that the whole of the available e.m.f. from the aerial is applied to the amplifier.

The pre-amplifier is connected to the main receiver through a long cable, and its output impedance should be low; a cathode-follower output stage is suitable.

4.5. *Summary of Requirements*

(a) High sensitivity (i.e., low inherent noise).
(b) Linearity of amplification.
(c) Overall voltage gain not greatly exceeding 2 at all frequencies between 2·5 and 20 Mc/s and very low gain at frequencies outside this band.
(d) Means for injecting a locally-generated signal into the first stage.
(e) High input impedance compared with that of the aerial.
(f) Output impedance approximately equal to that of the cable through which connection to the main receiver is made.

5. Design of the Pre-Amplifier

5.1. *Input Stage*

It has been shown[2] that a low inherent noise level can be achieved by means of a series connection of two valves in what has been called the 'cascode' circuit (see Fig. 2). The input valve V_1 is an earthed-cathode triode, the anode load of which is an earthed-grid triode, V_2, the same current passing through both valves. This combination has the high amplification of a pentode with the low noise of a triode, since it can be shown that the contribution of V_2 to the noise level is very small. The amplification of V_1 is of order unity, giving high stability and avoiding a large increase in the input capacitance of the valve due to Miller effect. All the gain is provided by V_2.

The noise level is almost entirely shot noise in V_1, provided that the grid-circuit impedance is low. Experiments were made with several types of valve and the 6AC7, used as a triode, was found to have a suitable combination of high mutual conductance and low noise level. The shot-noise level of this valve, expressed as an equivalent voltage on the grid, is rather less than 0·2 μV.

The anode load of V_2 is a network designed to provide uniform gain in the required band. The voltage gain is about 5 from the grid of V_1 to the grid of V_3, which is just sufficient to ensure that later stages do not contribute a large part of the total noise.

5.2. *Amplifier Stage*

It is necessary to raise the level of the noise of the first stage so that it predominates over the noise of the main receiver. The output stage is a cathode-follower in which some loss occurs, so an intermediate amplifier stage is required. This could be a pentode, as its noise characteristics are not of great importance, but the 6AC7 valve used is normally triode-connected as a convenient means of limiting the gain to the required value. The anode load is a network similar to that between the first two stages, and a further voltage gain between two and three is obtained.

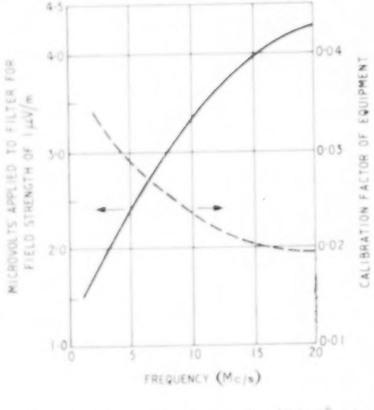


Fig. 3. *Pick-up factor of aerial and calibration factor of equipment; aerial length 6·1 m.*

5.3. *Output Stage*

The output stage is designed to feed a 70-ohm cable. It is in this stage that intermodulation is most likely to occur, since the signals at the grid have relatively high amplitude and the load impedance is small. A field of strength 10 mV/m at 5 Mc/s, for instance, produces a signal of about half a volt on the grid of V_4. Careful design is therefore necessary to avoid distortion.

A 6AC7 used as a cathode-follower was found to be insufficiently linear. Various other circuits were tried, including combination of valves with negative feedback, but the most satisfactory arrangement was obtained by using a 6AG7 valve as a simple cathode-follower, conditions being adjusted to give the most linear operation possible. Similar performance has been obtained by the use of a 6L6 valve. With a 70-ohm load there is a gain of $\frac{1}{4}$ in this stage.

5.4. *Aerial Circuit*

The aerial and its method of coupling to the grid of the first valve must be arranged to provide high sensitivity without introducing critical dependence of the pick-up factor on frequency. The effective height of the aerial itself should be as large as possible at the lowest frequencies without approaching the half-wavelength at the highest frequency, and the aerial length used with the original equipment (6·1 m) fulfils these requirements. Any increase in length, while improving the sensitivity, makes the calibration of the equipment markedly dependent on aerial properties and on ground constants, which cannot be accurately determined. For general use at outlying stations, therefore, the 6·1-m aerial is more satisfactory, but a longer aerial (7·6 m) has been used at one station (Tatsfield, England) where the noise level is so low that the additional sensitivity is especially valuable.

To reduce the effects of intermodulation a high-pass filter is connected between the aerial and the grid of V_1. It consists of a single T section with *m*-derived end sections, and by attenuating signals in the medium-frequency broadcast band by more than 40 db, greatly reduces the number of spurious signals generated in the pre-amplifier. The filter should preferably have an impedance large compared with that of the aerial as stated in Section 4.4, but the grid impedance of the valve is only a few hundred ohms at 20 Mc/s and a high-impedance filter cannot be properly terminated at this frequency. As a compromise the filter is designed for a load of 500 ohms which is roughly equal to the maximum aerial impedance. By this means the critical dependence of the pick-up factor on the aerial characteristics which was a weakness of the old arrangement (input impedance 60 ohms) is avoided.

The voltage applied to the filter for a field intensity of 1 μV/m has been calculated and plotted in Fig. 3 (solid curve) as a function of frequency. Field tests previously reported[1] have shown that the assumed aerial characteristics are substantially correct. The figure shows that there are no rapid variations which would make the calibration of the equipment critically dependent on frequency.

5.5. *Injection of the Artificial Signal*

The keyed signal from the generator is applied to a resistor of 4 ohms in the cathode of V_1, through a cable of impedance 70 ohms, a 65-ohm resistor being added in series to provide correct termination. This method of injection avoids the disadvantages of the old equipment mentioned in Section 2 (b and c). The voltage applied between cathode and grid of V_1 corresponding to a voltage *e* from the signal generator should be 0·06*e*, and the measured values are within ± 10% of this figure up to 10 Mc/s with a slight increase at higher frequencies.

5.6. *The Complete Circuit*

Fig. 2 shows the essential features of the pre-amplifier and the method of connection to the main receiver and signal generator. A resistor of 110 ohms is connected across the receiver to match the cable but, because the cable is not a part of the circuit in which the noise and artificial signal are compared, correct matching is not vital to the correct calibration of the equipment as it was with the original arrangement. A 70-ohm resistor is connected across the signal generator output in parallel with the cable to provide a correct match for the 35-ohm generator impedance, and the signal is keyed by a relay which disconnects the cable and substitutes a second 70-ohm resistor.

5.7. *General Arrangement of the Equipment*

The pre-amplifier unit is designed to be mounted on a concrete base some distance (normally 60 m) from the rest of the equipment, which incorporates the same units as are shown in Fig. 1. The mechanical arrangement of the unit is shown in Fig. 4. An annular casting is bolted to the concrete base and supports a second circular casting on which the pre-amplifier chassis is mounted. A steel, bell-shaped cover fits over the chassis and castings and supports the aerial, a tubular mast in three sections, mounted on an insulator at the top. A plug on the underside of the insulator engages with the input socket of the pre-amplifier as the cover is placed in position.

Radio-frequency coaxial cables connect the pre-amplifier to the main receiver and to the signal generator. A three-core, lead-covered cable supplies power to the unit and the valve heaters may be left permanently on to prevent condensation, the h.t. supply being switched independently.

6. Overall Performance of the Pre-Amplifier

The most important electrical characteristics of the pre-amplifier are the inherent noise level, the gain, the calibration factor and the linearity.

6.1. *Noise Level*

The set noise originates almost entirely in the first valve and its r.m.s. value, expressed as a voltage at the first grid, is rather less than 0·2 μV. The measured values referred to the aerial terminal are somewhat higher, due to a small loss in the filter (up to 3 db), and vary between 0·2 and 0·3 μV over the frequency band. Otherwise expressed, the noise factor of the receiver, used with a source of resistive impedance 500 ohms for which the receiver was designed, is about 3. However, since the aerial resistance varies with frequency between a few ohms and 1,000 ohms the noise factor referred to this resistance is variable and much higher than 3 at some frequencies.

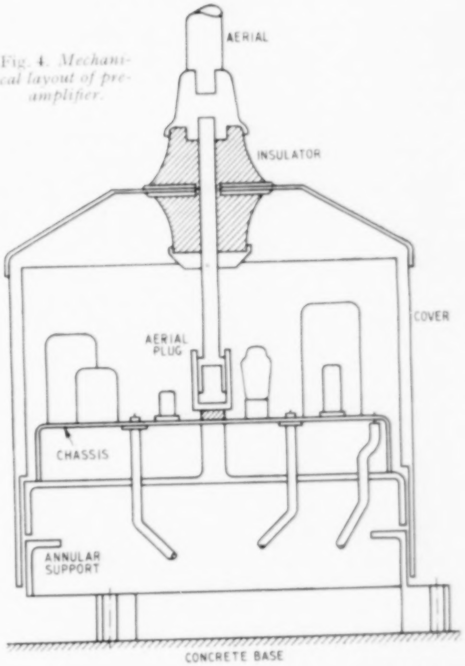


Fig. 4. *Mechanical layout of pre-amplifier.*

By considering the internal noise level in conjunction with the pick-up factor of the 6·1 m aerial (Fig. 3) it is found that the noise field strength equivalent to the set noise lies between 0·12 μV/m at 2·5 Mc/s and 0·07 μV/m at 20 Mc/s. With the 7·6 m aerial the figure at 20 Mc/s is reduced to 0·02 μV/m (at the expense of critical dependence on frequency). These figures represent approximately the lowest measurable noise fields, expressed as the c.w. field for intelligible reception through the noise, in a bandwidth of 10 kc/s.

6.2. *Gain*

The voltage gain of the pre-amplifier from the grid of the first valve to the output terminals is about 3 at all frequencies in the band, falling rather sharply above 20 Mc/s. This is sufficient to ensure that the main receiver contributes little to the total internal-noise level of the equipment. The effect of the filter is to introduce a sharp cut-off at 2 Mc/s, the attenuation of signals over most of the medium frequency range being greater than 40 db, and also to introduce a loss of up to 3 db in the pass band. The overall-gain characteristic is shown in Fig. 5.

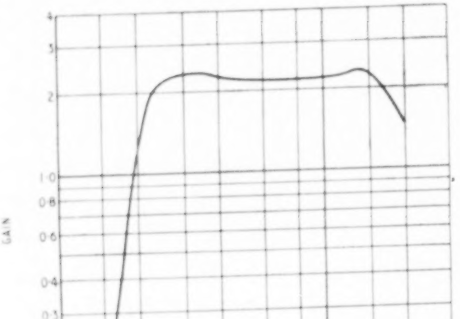


Fig. 5. *Gain characteristic of pre-amplifier.*

6.3. *Calibration Factor*

The voltage applied to the filter of the input stage for a given field strength has been discussed in Section 5.4 and is plotted in Fig. 3 (solid curve). The voltage applied to the cathode of the same stage for a given signal generator voltage e has been measured and found to increase from 0·06 at 2·5 Mc/s to about 0·08 at 20 Mc/s. The overall calibration factor, used to convert the signal-generator output to the equivalent field strength at the aerial is derived by combining the two above factors and making allowance for the small loss in the filter. The values are plotted in Fig. 3 (broken curve).

6.4. *Linearity*

The effects of non-linear amplification on the performance of the equipment have already been discussed, one effect being that two strong signals beat together and produce spurious signals at the sum and difference frequencies. It has been found that the amplitude of a spurious signal depends mainly on the product of the amplitudes of the two applied signals and not on their individual amplitudes. A useful measure of the degree of non-linearity is the product of the two applied signal amplitudes necessary to produce a spurious signal just audible through set noise.

It is found that when signals are applied to the pre-amplifier at two frequencies within the pass-band, the product of their amplitudes in mV must exceed 8 before an intermodulation signal can be heard. Taking the effective height of the aerial as 4 m, two equal signals must therefore each have a field strength of 700 μV/m to produce an audible intermodulation product. If one or both signals are below the pass band they must be some 40 db greater to produce the same interference.

It is not easy to measure the amplitudes of the harmonics generated by non-linear amplification of strong signals, because the output from commercial signal generators already has considerable harmonic content. Harmonics would be expected to attain significant amplitude at about the same input signal levels which produce appreciable intermodulation, and observations on strong broadcast transmissions indicate that this is so.

In view of the results of intermodulation tests with strong signals it is considered that the increase in apparent noise level due to intermodulation involving noise components is negligible. This view is confirmed by the fact that the measured intensity of noise arising in the first stage, in the presence of strong signals picked up on the aerial, is in agreement with the calculated value.

6.5. *Field Tests on the Complete Equipment*

One pre-amplifier has had extensive field tests, particular attention being paid to the sensitivity and to the effects of intermodulation. Pre-amplifiers have also been used for routine observations for two years at Tatsfield, England, and some months at other stations overseas.

To obtain a measure of the intermodulation and its effect on the difficulty of making measurements it was decided to estimate the number of spurious signals generated within certain defined frequency bands. The AR88 receiver, first connected directly to the aerial (with no cable), was operated with the beat oscillator on, and was tuned slowly through the band. The operator counted the number of heterodyne beats, great care being taken to include all those just audible through the noise. This process was then repeated immediately with the pre-amplifier in circuit. Assuming that the number of real signals was unchanged the difference in the number of beats heard was due partly to slightly increased sensitivity and partly to the generation of spurious signals, this being the main effect. The test was repeated several times, in a period of a few minutes, with similar results, so it was assumed that the number of real signals remained substantially constant.

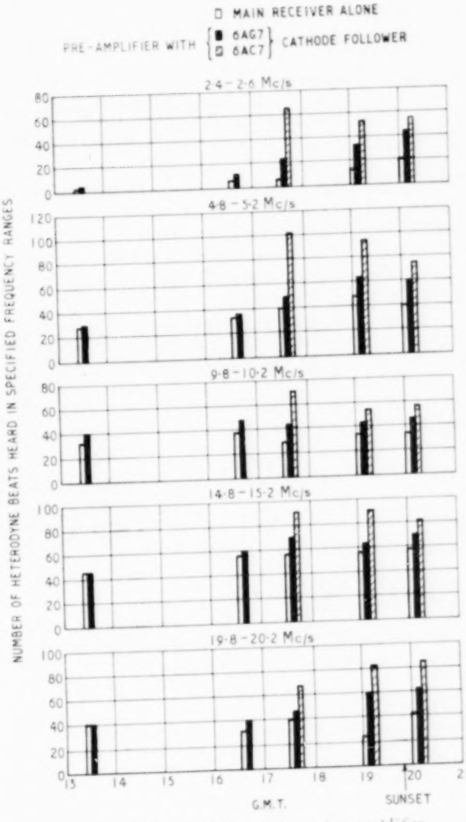


Fig. 6. *Intermodulation tests on pre-amplifier.*

The results are shown in Fig. 6, covering a period of several hours up to and just beyond sunset. The unshaded columns indicate the number of signals heard with the AR88 only and the blacked-out columns the number with the pre-amplifier. As sunset approached, the number of beats heard with the pre-amplifier tended to increase more rapidly than those heard with the main receiver alone, particularly at the lowest frequencies. Remembering that some of the increase in the number of signals was no doubt due to the slightly higher sensitivity of the pre-amplifier, it seems that at the worst the number of spurious signals is roughly equal to the number of real signals, and in general is much smaller. These tests were carried out with a pre-amplifier with higher gain (5) than that finally adopted (2); the performance of the final equipment should therefore be correspondingly better. Although any aggravation of the already difficult task of finding clear channels for noise measurements is unwelcome, it is considered that the increase in the number of interfering signals is tolerable and justified by the increased sensitivity of the equipment.

For comparison, some results obtained with a pre-amplifier with a 6AC7 output valve are shown (shaded). The inferior performance of this valve is evident.

The conclusions reached as a result of these listening tests have been borne out in the routine measurements. At Tatsfield, where many strong signals in the h.f. band are present and where noise levels are normally quite low, the use of a pre-amplifier has enabled measurements to be made of noise levels some 20 db lower than was possible previously, and the number of observations missed due to inability to find a clear channel has been quite small. At other stations where the pre-amplifier is in use noise levels are

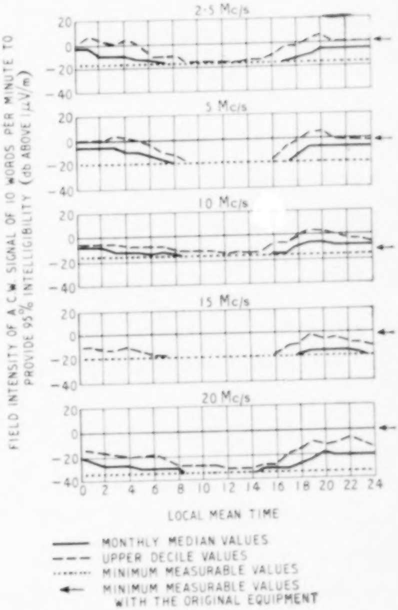


Fig. 7. *Measured noise levels at Tatsfield, England, for September* 1948.

in general higher and interference from stations no greater, and probably less than at Tatsfield. Typical noise curves taken at Tatsfield after installing the pre-amplifier are shown in Fig. 7 in which are plotted the monthly median values and the values exceeded by 10% of the readings for the month (upper decile values). With the original equipment the atmospheric-noise level was below the set-noise level for a major part of the time at the lower frequencies and for practically the whole time at the higher frequencies. The arrows, indicating the minimum measurable values with the original equipment, show that few readings would have been obtained during the month to which the plotted curves refer.

7. Conclusions

The wideband pre-amplifier described has made it possible to measure noise levels down to about 0·1 μV/m at the expense of some increase in the difficulty of taking the readings, as compared with the old system, due to intermodulation effects. A further advantage is that the calibration factor is substantially independent of the frequency, the ground constants and the characteristics of the receiving equipment other than the first stage of the pre-amplifier. In particular the cable connecting the pre-amplifier to the main receiver need not be precisely matched. The pre-amplifier was developed to enable a noise survey to be made on several frequencies between 2·5 and 20 Mc/s and it is considered that very little further improvement is possible without considerable elaboration in a system covering this band of frequencies. If the results of a further survey, using this equipment, indicate the need for measurements of even lower noise levels it will probably be necessary to concentrate on one frequency at a time and to design the equipment for maximum sensitivity at this frequency. This technique would slow down the rate at which data could be obtained, and is not justified at present.

8. Acknowledgments

The conception of a wideband pre-amplifier as a means of improving the sensitivity of the noise-measuring equipment was due to Dr. H. A. Thomas, who was also responsible for much of the design, in collaboration with staff of the National Bureau of Standards, Washington, D.C. The work of Messrs. R. Bateman, H. V. Cottony and P. Viezbicke of that organization is acknowledged. The design was completed and the test work carried out by the author.

The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of the National Physical Laboratory, and the Director of Radio Research of the Department of Scientific and Industrial Research.

REFERENCES

[1] Thomas, H. A. "A Subjective Method of Measuring Radio Noise." *Proc. Instn elect. Engrs*, 1950, Vol. 97, Part III, pp. 329–334.

[2] Wallman, H., MacNee, A. B., and Gadsden, C. P., "A Low Noise Amplifier." *Proc. Inst. Radio Engrs*, June 1948, Vol. 36, No. 6, pp. 700–708.

NEW BOOKS

Television Principles

By Robert B. Dome. Pp. 291 + xii. McGraw-Hill Publishing Co., Ltd., Aldwych House, London, W.C.2. Price 42s. 6d.

This book is of American origin and so deals primarily with television according to American standards. While this does not affect principles, it does show up in the practical examples which are chosen to illustrate them.

The book starts with an historical chapter and this is followed by one on electronic methods of scanning and reproduction. Video-frequency amplifiers are then treated, followed by transmitters, receiving and transmitting aerials, r.f. input circuits, i.f. amplifiers, detectors and scanning, in that order. The concluding chapters are a miscellaneous one and a discussion of propagation and relay systems.

The chapter on scanning is quite inadequate. It consists of 41 pages and includes a description of interlacing, the U.S. standards of sync-pulse waveform, methods of generating saw-tooth voltage waves, amplifiers for magnetic deflection, Fourier analysis of a saw-tooth, amplifiers for magnetic deflection; differentiators and integrators for sync-pulse separation and a.f.c. sync circuits. Most of these items are of sufficient importance to deserve chapters to themselves. There is, too, a great, lack of balance. Integrators occupy about seven pages, harmonic analysis about six, yet vertical sweep linearity is dismissed in eight *lines* and the modern efficiency-diode line-scan circuit has only some six pages of superficial treatment. Although shown in a circuit diagram, flyback e.h.t. is not mentioned in the text.

The earlier part of the book is much more satisfactory. The chapter on i.f. amplifiers, for instance, deals with i.f. coupling circuits and gives the formulae specifying their characteristics. Both coupled-circuit and stagger-tuned amplifiers are dealt with. The use of rejector circuits is considered briefly. On the r.f. side, noise is given pride of place, as it should be and, a rather surprising thing in an American book, very little space is given to station-selection systems. The section on frequency-changing is rather weak, occupying some four pages only in which there is no mention of oscillator frequency stability.

There is a surprising statement in the chapter on video amplifiers. After saying that the sharper the h.f. cut-off the greater are the overshoot and oscillation, the author goes on to say that "the oscillation really represents the negative of the missing Fourier components not transmitted". He then says that "it would seem desirable to design the first few stages of an amplifier chain with gradual cut-off characteristics and to design the last stages with sharp cut-off circuits, wherein the steep portion of the curve occurs in the attenuation band of the gradual cut-off stages". It is a plausible, but fallacious, argument; the response of a linear amplifier is dependent on the overall frequency and phase characteristics and the order of individual coupling networks of different characteristics in no way affects them.

The book as a whole is nicely produced and contains a great deal of useful information at a level suitable for the newcomer to television. Its value, however, is greatly reduced by the cursory treatment of scanning problems.

W. T. C.

Foundations of Wireless (5th Edition)

By M. G. Scroggie, B.Sc., M.I.E.E. Pp. 328, with 236 illustrations. Published for *Wireless World* by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 12s. 6d. (postage 8d.).

The 21 chapters of this book cover in an elementary manner all aspects of wireless from the nature of sound waves to transmission lines. Fundamental electrical ideas are clearly explained and the reader is taken through all the essential parts of receiving and transmitting equipment. The treatment is in the main explanatory; elementary mathematics are used where necessary but are never obtrusive.

F.B.I. Register of Manufacturers, 1951/52. (24th Edition)

Pp. 882. Published for the Federation of British Industries by Kelly's Directories, Ltd., and Iliffe & Sons, Ltd. Price 42s. (including postage.).

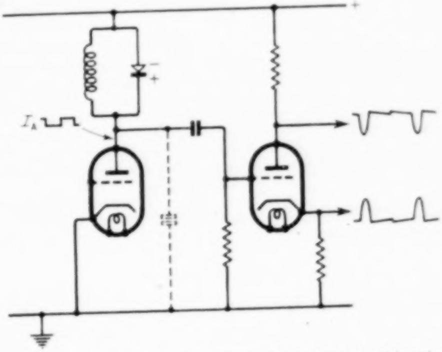
The only authorized directory of the Federation of British Industries. It includes Products and Services, Advertisements, Addresses, Trade Associations, Brands, Trade Names and Trade Marks. The information is classified for quick reference in English, French and Spanish.

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

R-C Coupling Network for Pulse Transmission

Sir,—Does not Dr. Hilary Moss's paper in your November issue serve chiefly to emphasize that the so-called differentiating circuit is a pretty poor method of getting sharp pulses from a square wave? I suggest that a low-capacitance ringing choke and rectifier combination, such as that shown in the figure, will give a better output pulse in nearly all cases.



E. F. Good.

Malvern, Worcestershire.
8th November, 1951.

Power Meter and Mismatch Indicator

Sir,—While the "Power Meter and Mismatch Indicator" described by Mr. Boff in the September issue is admittedly simple, it appears that several minor alterations in design would greatly enlarge its scope and increase its accuracy.

The sudden change in section of the coaxial line shown diagrammatically in Fig. 5(b) and confirmed in Fig. 3, can hardly fail to produce reflections causing inaccuracy according to frequency and termination conditions. Pye-type connectors of the normal design will also tend to introduce some reflection.

The paper on Power Measurement in *J. Instn elect. Engrs*, Vol. 93, Part IIIA, p. 1455, by Crowley-Milling, Gordon, Miller and Saxon, describes a feed-through wattmeter for 500-600 Mc/s, employing a simple directional-coupling loop after the system proposed by Pistolkors and Neumann in *Elektrosvyaz*, 1941, Vol. 9, No. 4. This system, although undoubtedly more frequency conscious than the short loop, was found to give reasonably fair results. The use of bolometer or vacuo-junction detectors which give a closer approximation to square-law (with some loss of sensitivity), and the advantages of electrical subtraction of the readings of direct and reflected power are worth considering.

Some figures indicating the accuracy of the device for various standing-wave ratios over the frequency range would be of interest, particularly as it appears that the values obtained for direct, reflected and, hence, net transmitted power, will depend in a complex manner on frequency, s.w.r. and phase of the reflected wave.

It is agreed that the directional power meter is a simple and robust instrument of manifold uses which may be directly calibrated and quite accurate over a limited frequency band.

Douglas S. Gordon.

Electrical Engineering Dept.,
The University, Glasgow.
18th November, 1951.

Absolute Measurement of Phase Shift

Sir,—After reading Mr. Wintle's paper in your July issue, and examining the references quoted together with a number of others, I am led to suppose that the technique to be described here is not well known and may therefore be of general interest. It is an absolute method in the sense that the calibration involves only a process of counting.

The equipment required is not complex and will generally be available in a research laboratory. The precision of phase measurement is high, being of the order of ± 0·25° (independent of the magnitude of the angle), even under relatively unfavourable conditions. The principle is simple, and measurements are made much more rapidly than a somewhat lengthy description might suggest.

Condensed to its bare essentials the process is as follows: a nine-to-one Lissajous figure is produced on a c.r.-oscilloscope screen, the X-plates being fed from the 'master' oscillator (frequency f_1) through the circuit under examination. The Y-plates are fed from a 'slave' oscillator, running at a frequency 9 f_1, and locked to the 'master' oscillator by a spike signal. This spike can be shifted in phase relative to the voltage under examination, by means of one or other of two phase shifters, P_1 and P_2. P_1 is not calibrated and has a fairly wide range of phase shift (> 180°). P_2 is calibrated (from the equipment itself) over a range of 10 or 20 degrees.

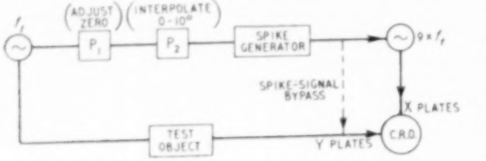
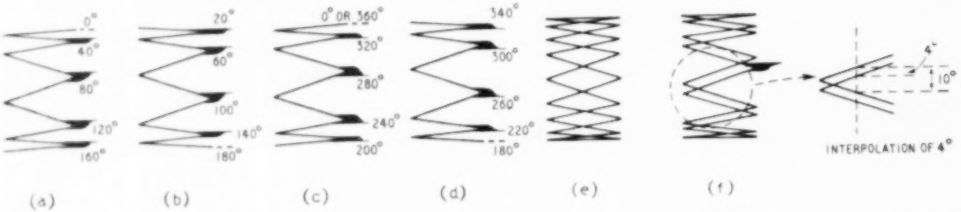


Fig. 1. (above). Fig. 2. (below).



If either P_1 or P_2 is varied the c.r.o. trace changes through the various phases illustrated in the sequence 2(f), 2(e), 2(a), 2(e), 2(b), 2(e), etc. The change from the serpentine 2(a) to the symmetrically-looped figure 2(e) occurs at intervals of 10° phase shift. The change from 2(a) to 2(b) occurs at 20° intervals. In 2(f), a spike is shown superimposed on the figure. This is obtained by by-passing part of the locking spike on to the X-plates.

The position of the spike on the trace is an indication of phase shift. An ambiguity of 180° can be resolved if the spike is made slightly asymmetrical on the time scale.

A set of reference diagrams 2(a)-2(d) can be drawn up showing the position of the spike corresponding to every 20° of phase-shift reckoned from an arbitrary zero. Using these diagrams a trace such as Fig. 2(f) indicates 84°.

Interpolation between the 20° points can be made either by eye, or with precision, by the use of the phase shifter P_2. This can be calibrated by a technique similar to that just described but using a frequency ratio other than 9/1. For instance we have:

Frequency Ratio	Phase shift corresponding to:—	
	Looped Figure	*Serpentine Figure*
10	9°	18°
12	7½°	15°
15	6°	12°
18	5°	10°

Using these degree intervals as a guide, a calibration curve of P_2 can be drawn which can be relied upon at the interpolated points, to an accuracy better than ± 0·25°.

It will be appreciated that the degree markings in Figs. 2(a)-2(d) have a purely arbitrary origin. The sign is arbitrary also and must be checked by measuring the phase shift through, for instance, an elementary RC network.

As described, the apparatus will measure the difference of phase between any two points in a network: to do this the 'marking' spike is needed. But when the apparatus is used to calibrate a phase-shifter at 10° intervals the spike and the diagrams 2(a)-2(d) are not required. It is necessary then only to count the number of reversals of the serpentine figure and multiply by 20°, and to interpolate the 10° by the use of the looped figure.

For general purposes, where a precision of the order of ± 1° is considered good enough, the phase shifter P_2 may be dispensed with and the fraction of 10° estimated from the figure by eye.

For measurements of the highest precision the waveform of the master oscillator should be good, otherwise the serpentine figures do not close up to a clean line and the looped figure is not symmetrical. Tests show that errors introduced by 2-3% of second and third harmonics appear to have a standard deviation of the order of 0·25°.

A 2-valve zero-phase shift RC oscillator serves very well as a 'slave'. The spike voltage derived from the 'master' is injected across a resistance in the common h.t. lead to the valve anodes. No great stability is required as the spike signal readily holds the 'slave' in synchronism.

If the locking signal is made too powerful, the Lissajous figures are distorted and do not show a clean line. The position of the spike on the c.r.o. trace can be adjusted by detuning the slave oscillator.

E. R. WIGAN.

Barnham, Sussex.
12th November, 1951.

Poynting's Theorem

SIR,—I have just come across your Editorial on this Theorem in your September 1951 issue.

It is not generally known that Heaviside developed this same theorem independently though somewhat later than Poynting. Heaviside expressed the result for plane waves in a slightly different form. According to Heaviside the divergent power vector $\mathbf{E} \times \mathbf{H}$ becomes $v(U + T)$; where U and T are the electric and magnetic energy densities and v the velocity of the wave.

Heaviside's expression gives a far more tangible interpretation of Maxwell's theory and will explain at once why there is no continuous flow of power in the apparently anomalous case you cite and illustrate.

G. A. M. HYDE.

Kirkby,
Nr. Liverpool.
14th November, 1951.

ERRATUM

In the paper "Mutual Impedance of Wire Aerials", by L. Lewin, in the December 1951 issue, p. 352, the last term in equation (17) should read 2 log (cot ½ θ) not 2 log (cos ½ θ).

ABSTRACTS and REFERENCES

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research and published by arrangement with that Department.

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

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ACOUSTICS AND AUDIO FREQUENCIES

534.2 : 551.556 1

Wind Noise and the Transmission of Sound in the Open Air.—F. Spandöck. (*Z. angew. Phys.*, June 1951, Vol. 3, No. 6, pp. 228–231.)

534.231 2

The Radiation of Pulses by Plane Piston Diaphragms in a Rigid Wall.—F. A. Fischer. (*Acustica*, 1951, Vol. 1, No. 1, pp. 35–39. In German.) When pulses are radiated by a piston diaphragm, direction-dependent distortion occurs because the propagation of the various spectral components varies in accordance with the frequency dependence of the radiation pattern. Calculation shows that a radiated unit pulse can be considered as composed of individual pulses from elementary areas of the diaphragm, and gives a representation of the shape of the diaphragm. For a pulse of arbitrary shape the resultant is a product depending on both diaphragm shape and pulse shape.

534.231 + 621.396.67] : 778.3 3

A Photographic Method for Displaying Sound-Wave and Microwave Space Patterns.—W. E. Kock & F. K. Harvey. (*Bell Syst. tech. J.*, July 1951, Vol. 30, No. 3, pp. 564–587.) Description of the method, with many illustrations of actual sound-wave and microwave fields. A similar probe method, for recording on Teledeltos paper, has been described by Iams (3590 of 1947).

534.232 : 534.321.9 4

Construction of High-Power Ultrasonic Transmitters.—E. Skudrzyk. (*Elektrotech. u. Maschinenb.*, 1st & 15th April 1951, Vol. 68, Nos. 7 & 8, pp. 173–178 & 202–212.) Discussion of both quartz and magnetostriction oscillators. Suitable types of holder for high-power quartz oscillators are described and an oscillator transmitting waves through a duralumin membrane, for therapy applications, is illustrated. Simple equivalent-circuit theory of quartz oscillators is outlined and more rigorous theory is considered. Equivalent-circuit theory is also given for magnetostriction oscillators and a numerical example is calculated. Methods of measuring the output of ultrasonic oscillators are indicated.

534.232 : 534.321.9 5

An Efficient Low-Power Ultrasonic Generator.—F. Pirker. (*Radio Tech., Vienna*, April 1951, Vol. 27, No. 4, pp. 175–180.) Description of the construction and circuit arrangement of a generator which provides an output of 50 W at an electroacoustic efficiency of 60%. Full details are given of the preparation and mounting of the quartz crystal, which is energized from a Hartley circuit using a 40-W pentode.

534.321.9 6

Ultrasonics in Air, and Applications.—F. Canac & V. Gavreau. (*Acustica*, 1951, Vol. 1, No. 1, pp. 2–16. In French.) Electrodynamic and magnetostrictive generators and receivers are described, and investigations of the reflection and diffusion of ultrasonic waves by cylinders and polycylindrical surfaces are reported. Practical acoustical applications discussed include the study of hall acoustics by means of models.

534.321.9 : 061.3 7

Ultrasonics in Fluids.—E. G. Richardson. (*Nature, Lond.*, 21st July 1951, Vol. 168, No. 4264; pp. 106–107.) Brief report of an international conference held in Brussels in June 1951. About 35 papers were presented, dealing with physical measurements and theories in the field of ultrasonics.

534.321.9 : 534.373] : 538.221 8

The Influence of Magnetization on Ultrasonic Attenuation in a Single Crystal of Nickel or Iron-Silicon.—Levy & Truell. (See 169.)

534.44 9

An 8 000-c/s Sound Spectrograph.—O. Gruenz. (*Bell Lab. Rec.*, June 1951, Vol. 29, No. 6, pp. 256–261.) Spectral density in 45-c/s and 300-c/s bandwidths at frequencies up to 8 kc/s is permanently recorded on charts. Energy distribution with frequency is presented in two ways: (*a*) as discrete horizontal markings corresponding to different frequencies, for any selected 5-ms period of integration, with a maximum of 35 db relative amplitude; (*b*) as a continuous record of intensity, indicated by relative darkness of trace, in a frequency/time rectangular-coordinate framework.

534.75 **10**

Nonlinear Characteristics of the Ear.—G. Haar. (*Funk u. Ton*, May 1951, Vol. 5, No. 5, pp. 248–257.) Comparative tests of the response of the ear to two pure notes of different frequencies f_1 and f_2 received simultaneously show that difference tones formed in the ear are physically real vibrations which occur even when they are inaudible. The predominant difference frequency is $2f_1 - f_2$.

534.833.1 **11**

A Tentative Method for the Measurement of Indirect Sound Transmission in Buildings.—E. Meyer, P. H. Parkin, H. Oberst & H. J. Purkis. (*Acustica*, 1951, Vol. 1, No. 1, pp. 17–28. In English.)

534.845.2 **12**

Acoustic Behaviour of a Porous Material.—A. Bressi & G. G. Sacerdote. (*Alta Frequenza*, Feb. 1951, Vol. 20, No. 1, pp. 28–33.) Measurements are reported of the acoustic absorption of 'betamianto'. The results differ from those for other porous materials in that an absorption maximum is exhibited at low frequencies even for a thin layer.

534.846 **13**

Acoustics and Sound Exclusion.—W. A. Allen & P. H. Parkin. (*Archit. Rev., Lond.*, June 1951, Vol. 109, No. 654, pp. 377–384.) Survey of the planning, structure and finishing materials of the Royal Festival Hall, London, with consideration of both the internal acoustics of the auditorium and the exclusion of noise from the outside.

534.846.6 **14**

Investigation of Sound Diffusion in Rooms by means of a Model.—T. Somerville & F. L. Ward. (*Acustica*, 1951, Vol. 1, No. 1, pp. 40–48. In English.) An experimental comparison was made of the effects produced by rectangular-, semicylindrical- and triangular-section diffusing elements applied to the room walls; the investigation covered both steady-state and pulse operation. All three types of diffuser caused reduction of irregularity in both types of characteristic, the effect produced by the rectangular-section diffusers being greatest.

621.395.61/.62 **15**

Standards on Transducers: Definitions of Terms, 1951.—(*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 897–899.) Reprints of this Standard, 51 IRE 20 S2, may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, N.Y., at $0·50 per copy.

621.395.61/.62 : 621.395.92 **16**

Acoustic Transducers for Hearing Aids.—W. Güttner. (*Fernmeldetech. Z.*, May 1951, Vol. 4, No. 5, pp. 227–234.) A review of piezoelectric transducers. For the production of crystal microphones the most suitable materials are Rochelle-salt and ammonium-phosphate crystals and polarized $BaTiO_3$ ceramic. The properties of such microphones are analysed and their parameters evaluated. For crystal ear-pieces using Rochelle salt, all possible constructions are described. From their mechanical equivalent representations the frequency characteristics of the sound pressure are calculated and compared with measurement data.

621.395.61 : 621.385.82.029.3 **17**

Increasing the Efficiency of the High-Power Thermionic Cell by Superposition of a Strong Field obtained from a High Voltage of High Frequency.—Klein. (See 288.)

621.395.616 **18**

Electrical Input Resistance of Capacitor Microphone.—U. Kirschner. (*Arch. elekt. Übertragung*, June 1951, Vol. 5, No. 6, pp. 273–278.) Calculation of the effective resistance is based on the work of Braun (1087 of 1945) and Weymann (470 of 1944). The results are applied to determine the parameters of a capacitor microphone to be used as a sound-pressure receiver with a circular characteristic.

621.395.623.7 **19**

Progress in the Development of Electrodynamic Loudspeakers.—G. Buchmann & K. Küpfmüller. (*Fernmeldetech. Z.*, June 1951, Vol. 4, No. 6, pp. 253–261.) The frequency response curve, the sensitivity and the nonlinear distortion give together a good idea of the quality of a loudspeaker. Suitable methods for determining these quantities are described. The optimum response curve is found to be one which falls off at both the lower and the higher frequencies. With suitable design, cheap loudspeakers can be produced with a good response curve over a wide band of frequencies.

621.395.623.7 **20**

Acoustic Boosting of the Lower Audio Frequencies by means of Bass Resonators with Phase Inversion.—H. Gemperle. (*Radio Tech., Vienna*, June 1951, Vol. 27, No. 6, pp. 245–249.) The method makes use of a resonator forming part of the loudspeaker housing and coupled acoustically to the back of the vibrating membrane. Optimum dimensions of the resonator are calculable for any particular loudspeaker. Practical tests showed an increase in loudness of frequencies in the range 30–100 c/s of about 8 db when the resonator was designed for a frequency of 60 c/s.

621.395.623.7 **21**

Acoustic Coupling of the Diaphragms in Duo-Cone Loudspeakers.—G. B. Madella. (*Alta Frequenza*, Oct./Dec. 1950, Vol. 19, Nos. 5/6, pp. 267–276.) Consideration of an idealized case indicates that radiated power as a function of diaphragm velocity is appreciably affected by the degree of coupling.

621.395.625 + 621.396.62] : 061.4 **22**

The Radio Exhibition at the Vienna Spring Fair.—(See 250.)

621.395.625.3 **23**

Recording Demagnetization in Magnetic Tape Recording.—O. W. Muckenhirn. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 891–897.) "An analysis of the magnetic tape recording process employing supersonic excitation is presented by considering the effect of the spatial distribution of the magnetic field around the recording head air gap on the magnetic history of an unmagnetized element of tape as it tracks across the recording head. This leads to an effect which is termed 'recording demagnetization' and serves to explain certain performance characteristics. An experimental technique developed for the measurement of this recording demagnetization is described, as is the method of measuring the air-gap field distribution. Finally, the correlation of the measurements of the recording demagnetization with normal recording performance characteristics is reported."

621.395.625.3 **24**

Magnetic Recording.—M. Alixant. (*Radio tech. Dig., Édn franç.*, 1951, Vol. 5, No. 3, pp. 147–161.) A bibliography of articles and patents complementary to that included in the book 'Magnetic Recording' by Schuh & Mikhnewitch published in 1950.

621.395.625.3 **25**

The R.C.A. 10794 High-Fidelity Magnetic Recording Head.—M. Rettinger. (*Radio tech. Dig., Édn franç.*, 1951, Vol. 5, No. 3, pp. 131–141.) French version of 807 of 1951.

621.396.645.029.4 : 621.3.018.78† : 534.861 **26**

Audio-Frequency Amplifiers with Adjustable Nonlinearity (Distortion Circuits).—G. Hoffmann. (*Funk u. Ton*, April 1951, Vol. 5, No. 4, pp. 169–176.) Two amplifier circuits are described having nonlinear characteristics largely independent of frequency and designed for investigation of the audibility of distortion tones produced in the transmission of music.

AERIALS AND TRANSMISSION LINES

621.315.2 **27**

Interaxial Spacing and Dielectric Constant of Pairs in Multipaired Cables.—J. T. Maupin. (*Bell Syst. tech. J.*, July 1951, Vol. 30, No. 3, pp. 652–667.) Experiments show that so far as the interaxial spacing and dielectric constant are concerned, actual cable conditions can be represented by two parameters of an ideal cable. These two parameters can thus be readily obtained for a commercial cable by calculation from the measured inductance, mutual capacitance and capacitance to ground of the pair.

621.315.2 : 621.396.97 **28**

Modern Broadcasting Cables to C.C.I.F. Recommendations.—E. A. Pavel. (*Fernmeldetech. Z.*, April 1951, Vol. 4, No. 4, pp. 150–157.) The essential features of the former and the recent C.C.I.F. recommendations are compared and tabulated, with particular reference to the widening of the transmission band required in connection with the introduction of u.s.w. broadcasting.

621.392 : 621.396.677 **29**

Approximation Methods in Radial Transmission Line Theory with Application to Horns.—A. E. Laemmel, N. Marcuvitz & A. A. Oliner. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 959–965.) "The radial transmission line theory provides a rigorous description of the propagation of electromagnetic energy in certain cylindrical regions. The complexity of direct calculation may be avoided by a simpler, although approximate, description based on the asymptotic identity of radial and uniform transmission line formulas at large radii. Approximate expressions are developed for input admittance, frequency sensitivity, and higher mode interaction effects on radial lines. Estimates of the order of accuracy are included, and applications are made to sectoral horns."

621.392.09 **30**

Propagation of an Electromagnetic Wave Guided by a Metal Surface with Dielectric Coating.—J. F. Colin. (*Onde élect.*, May 1951, Vol. 31, No. 290, pp. 245–256.) Propagation along a cylindrical conductor is considered, assuming a perfect conductor and a loss-free dielectric. The conditions are deduced for a single possible mode of propagation, the TM_{01} mode; if the frequency, or the dielectric constant, or the thickness of the dielectric are increased beyond certain limits, conditions are obtained in which other modes are possible. The effective cross-section of the conductor in transmitting power is much greater than its physical dimensions, owing to the distribution of energy in the surrounding fields. Expressions are deduced for the effect of losses and for the maximum power transmitted, and curves are drawn to assist in the solution of practical cases. The conditions for propagation of TE waves are briefly discussed.

621.392.2.09 + 621.396.11 **31**

Characteristic Impedance, Power, Voltage and Current in Transmission along Lines, in Waveguides and in Free Space.—O. Zinke. (*Funk u. Ton*, May 1951, Vol. 5, No. 5, pp. 225–238.) Neglecting losses and assuming purely progressive waves of sinusoidal form, the following relations are derived: (*a*) the ratio of the mean field strengths E_m/H_m at any point in the field cross-section perpendicular to the direction of propagation is independent of the dimensions of the system and is termed the elementary characteristic impedance z_e; (*b*) the characteristic impedance z_s of the system is equal to $z_e h_m/b_m$, where h_m is the effective height and b_m the effective breadth of the field; (*c*) the power transfer $N = E_m \cdot H_m \times h_m \cdot b_m$, i.e., the product of mean power per cm^2 and the effective cross-section S of the field. For rectangular waveguides S agrees exactly with the actual cross-section; for coaxial cables and narrow strip conductors the agreement is only approximate.

621.392.26† **32**

Experimental Investigation of the Reflections produced in a Waveguide by any Dielectric.—L. R. Noriega. (*Radio franç.*, June 1951, No. 6, pp. 1–9.) Reprint. See 3038 of 1949.

621.392.26† **33**

A Study of Single-Surface Corrugated Guides.—W. Rotman. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 952–959.) Two types of structure were investigated experimentally as transmission lines and as radiators: (*a*) a flat grooved plate fed by a waveguide; (*b*) a circular corrugated cylinder fed by a coaxial line. Modification of type (*b*) results in a spirally grooved rod with similar properties. Measured field parameters agree with those predicted from theory. For properly designed structures the energy is essentially bound to the corrugated surface; little radiation occurs, and the attenuation of the travelling wave is due mainly to losses in the metal. The effect of filling the corrugations with solid dielectric is analysed.

621.392.26† : 621.39.09 **34**

Refraction of Evanescent Waves.—N. Carrara. (*Alta Frequenza*, Aug. 1950, Vol. 19, No. 4, pp. 164–174.) An account of experiments in which evanescent centimetre e.m. waves are incident on a plane separating two dielectrics, the reflected wave being also evanescent, while the refracted wave is of ordinary type. The results indicate that Snell's laws and Fresnel's formulae are valid even when the angles of incidence and reflection are complex.

621.392.26† : 621.39.09 **35**

Propagation of U.H.F. Electromagnetic Waves in Conducting Waveguides of Rectangular Cross-Section: Part 1.—H. Frühauf. (*Elektrotechnik, Berlin*, June 1951, Vol. 5, No. 6, pp. 263–269.) Under certain conditions, energy is propagated in tubular conductors as plane waves. Vector analysis is applied to derive simple formulae, applicable by mathematical technicians to practical cases, for the field components of E and H waves, the limiting frequency, the wavelength in the rectangular waveguide, and the phase velocity.

621.396.67 **36**

Progressive-Wave Aerials.—A. F. Huerta. (*Rev. Telecomunicación, Madrid*, June 1951, Vol. 6, No. 24, pp. 23–37.) Theory is given relating to a progressive-wave aerial terminated by its characteristic impedance; the gain is calculated and the effect of finite ground conductivity is discussed. The results are applied to explain the properties of rhombic aerials and of the Marconi series-phase aerial.

621.396.67 **37**

A Short Proof that an Isotropic Antenna is Impossible.—H. F. Mathis. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, p. 970.)

621.396.67 **38**

Tuning and Matching of Aerials.—A. F. Huertas. (*Rev. Telecomunicación, Madrid*, March 1951, Vol. 6, No. 23, pp. 38–49.) Various methods of tuning and matching are illustrated and analysed. Particular cases considered include the excitation of a directive group of aerials and applications of 'balun' devices.

621.396.67 : 621.316.761.2 **39**

Increase of the Bandwidth of Aerials by means of Compensation Circuits.—R. Goublin. (*Onde élect.*, May 1951, Vol. 31, No. 290, pp. 233–244.) The problem is reduced to that of determining an aerial and associated network such that, in the complex diagram of the system, its input impedance falls within a circle defined by the magnitude of the maximum permissible s.w.r. in the feeder line. Using this construction the general cases are discussed of the matching of any aerial to a network and of the increase of aerial bandwidth by compensation circuits, or by sections of transmission lines. Self-compensated arrays are described and compensation methods are demonstrated by practical examples.

621.396.67 : 621.397.5 **40**

Directional Operation Planned for KRON-TV.—R. A. Isberg. (*Tele-Tech*, July 1951, Vol. 10, No. 7, pp. 26–27, 76.) An 8-section channel-4 R.C.A. TF-DA 'Supergain' aerial is mounted on a 200-ft tower on San Bruno Mountain, 4 miles inland from the San Francisco coast. Radiation is at present omnidirectional, but may be modified to give 100 kW radiated power inland and 3 kW radiated power towards the ocean. Feeder and aerial assembly are described in outline.

621.396.67 + 534.231] : 778.3 **41**

A Photographic Method for Displaying Sound-Wave and Microwave Space Patterns.—Kock & Harvey. (See 3.)

621.396.67.029.51 **42**

A Long-Wave Aerial.—L. C. Garcia. (*Rev. Telecomunicación, Madrid*, March 1951, Vol. 6, No. 23, pp. 82–85.) Description of an aerial forming part of the equipment of the new Spanish Transradio station at Parets del Vallés, Barcelona. The aerial is of the inverted-L type, the horizontal portion having a mean height of 86 m and being supported from steel lattice masts. The operating frequency of 79·15 kc/s will give a working radius of 1 750 km with fields > 30 μV/m, sufficient for reception on an undulator.

621.396.67.029.62 **43**

Transmitting Aerials for U.S.W. Broadcasting.—W. Berndt. (*Telefunken Ztg*, March 1951, Vol. 24, No. 90, pp. 6–21.) The characteristics desirable in aerials for broadcasting in the frequency band 87·7–100 Mc/s are enumerated and the properties of simple and beamed radiator elements are described, the conditions being determined for selection of the optimum distance between the elements. Methods of feeding multi-element aerials are described. Practical installations of U-aerials and double-slot aerials are illustrated and discussed.

621.396.67.029.63 **44**

Sidefire Helix U.H.F.-TV Transmitting Antenna.—L. O. Krause. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 107–109.) A travelling-wave helical aerial giving a power gain of 20 at 500 Mc/s, with a bandwidth of 20 Mc/s, consists of four vertically stacked bays, each including a right- and a left-hand helix placed end to end and fed at their junction. Beam tilt can be produced by mechanically rotating one portion of the aerial relative to the other.

621.396.67.029.64 **45**

U.H.F. Dipoles.—P. E. Vincelet. (*Rev. tech. Comp. franç. Thomson-Houston*, April 1951, No. 15, pp. 43–49.) In the construction of linear arrays dipoles, including slotted dipoles, are particularly useful. A simplified method of calculating the characteristics of dipole arrays is outlined and admittance characteristics of dipoles used under various conditions are given.

621.396.671 **46**

[E.m.] **Waves in a Flat Horn.**—B. L. Rozhdestvenski. (*C. R. Acad. Sci. U.R.S.S.*, 11th March 1951, Vol. 77, No. 2, pp. 221–224. In Russian.) A mathematical discussion is presented of the excitation of a flat horn by waves, either plane or cylindrical, arriving from a waveguide coupled to the horn. A conformal transformation is used which enables the problem to be reduced to a differential equation with partial derivatives. With certain simplifying assumptions a finite solution of this equation can be obtained. The solution is sufficiently accurate for practical purposes but its accuracy can be increased by using the method of successive approximations. The method is applicable also to circular and coaxial horns.

621.396.676 **47**

Radio-Frequency Current Distributions on Aircraft Structures.—J. V. N. Granger & T. Morita. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 932–938.) Models having scale factors of 43 : 1 to 72 : 1 were used, and the magnetic field surrounding the excited model was explored by means of a small loop. The results are shown diagrammatically for several types of aircraft excited by aerials of different types.

621.396.676 **48**

Experimental Study of Slot Aerials.—J. Desauty & R. Rocherolles. (*Radio franç.*, June 1951, No. 6, pp. 14–17.) Signals from a slot aerial located under an aircraft were received by a similar aerial on the ground. Polar diagrams are given; the results are in conformity with theory.

621.396.677 **49**

Lenses for Microwaves.—F. I. Garrido. (*Rev. Telecomunicación, Madrid*, March 1951, Vol. 6, No. 23, pp. 65–75.) A concise review of the properties of various types of lens, including those formed of homogeneous dielectric, those with plane-parallel conductors, lenses of artificial dielectric and others in which the index of refraction varies with distance from the axis.

621.396.677 **50**

Path-Length Microwave Lens.—D. G. Kiely. (*Wireless Engr*, Aug. 1951, Vol. 28, No. 335, pp. 248–250.) Experimental investigation of the E-plane radiation patterns of a cylindrical path-length lens shows an asymmetric main beam, which cannot be caused solely by an asymmetric distribution of field amplitude across the aperture but must be due also to the presence of asymmetric phase curvature, contrary to the experience of Kock (3058 of 1949). The path-length lens, although having many H-plane applications, is fundamentally unsuited for phase-correction in the E-plane.

621.396.679.4 **51**

New Method of Top Feed for Antifading Broadcasting Masts.—H. Graziadei. (*Fernmeldetech. Z.*, April 1951, Vol. 4, No. 4, pp. 159–167.) Methods adopted to reduce fading effects in the service area of a broadcasting aerial are discussed and the advantages of feeding an aerial at a point near the top are pointed out. Practical methods for top feeding a vertical aerial are described and the results of measurements of the vertical distribution of radiation for aerials fed at the foot and top respectively are shown in graphs.

61.392.26† **52**

Felder und Wellen in Hohlleitern. [Book Review]—H. H. Meinke. Publishers: R. Oldenburg, München, 1949, 148 pp., 15 DM. (*Frequenz*, Feb. 1951, Vol. 5, No. 2, p. 53.) A detailed exposition in relatively simple form of the propagation of waves in waveguides. Useful for practising engineers as well as for students.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.012.3 : 621.396.621.54 **53**

New Diagrams for 'Ganging' Calculation.—H. Kerbel. (*Funk u. Ton*, June 1951, Vol. 5, No. 6, pp. 287–296.) Diagrams facilitating the determination of the parameters of ganged capacitors in superheterodyne receivers are given. From the data of the variable capacitors completely independent parameters can be derived. The values of the parallel (trimmer) and series capacitors are obtained by multiplication of the corresponding parameters by the capacitance variation ΔC of the variable capacitor, the inductance by division by ΔC. The magnitude and distribution of the distributed capacitance are taken into account. A method of measuring the error curve is also described.

621.3.015.7† : 621.387.4† **54**

A Mechanical Kick-Sorter (Pulse Size Analyser).—S. G. F. Frank, O. R. Frisch & G. G. Scarrott. (*Phil. Mag.*, June 1951, Vol. 42, No. 329, pp. 603–611.) Each pulse causes a small steel ball to be propelled along an inclined board; the ball describes a parabolic path, which is determined by the pulse amplitude, and lands in one of 30 parallel grooves. A histogram of the pulse size distribution is thus built up. The mechanical construction and associated electronic circuits are described. The latter eliminate pulses which are either too large or too small to be recorded, or which follow too closely upon the previous pulse.

621.3.015.7† : 621.3.018.78† **55**

Pulse Distortion.—S. H. Moss. (*Proc. Instn elect. Engrs*, Part III, Sept. 1951, Vol. 98, No. 55, pp. 398–400.) Summary of I.E.E. Monograph No. 5. For a linear low-pass process, certain invariants which completely define the pulse shape exist for any input pulse; these are always increased by the same invariants, imposed by the system, to yield the invariants of the output waveform. The pulse invariants correspond to parameters used to describe statistical distributions. The theory in generalized form applies also to wave packets.

621.3.016.35 **56**

Application of Bode's Relation to the Calculation of a Nyquist Diagram.—A. Herrent & G. Novgorodsky. (*HF, Brussels*, 1951, No. 9, pp. 229–236, 255.) In applying Nyquist's stability criterion to a closed system it is necessary to know the frequency/phase and frequency/amplitude characteristics of the network over a very wide frequency range. The practical difficulties involved, especially at low frequencies, are eliminated by the use of Bode's relation which relates the phase and amplitude characteristics. A quick and easily applied graphical method is described for evaluating the integral in Bode's formula and hence the Nyquist diagram may be drawn. The method has been verified experimentally in the case of a low-frequency amplifier with resistive feedback.

621.314.2 : 621.392.6 **57**

Contributions to the Theory of the Differential Transformer.—G. Schmitt. (*Frequenz*, Feb. 1951, Vol. 5, No. 2, pp. 39–44.) Equations and equivalent circuits for the differential transformer are developed from the theory of six-pole networks. The stop-band attenuation is calculated and its frequency variation investigated. Design conditions are discussed for making the stop-band attenuation as great as possible.

621.315.592† : 537.312.6 **58**

Experiments on Thermistors of Italian Manufacture.—P. Lombardi & M. Vallauri. (*Alta Frequenza*, April 1951, Vol. 20, No. 2, pp. 51–67.) Basic theory of electronic conduction in semiconductors is outlined, and variation of resistance with temperature is considered. Measurements on specimen thermistors are reported, and the values of their time constants are deduced. Applications both familiar and unfamiliar are discussed.

621.316.726.078.3 **59**

U.H.F. Discriminator and its Application to Frequency Stabilization [of klystron].—G. Pircher. (*Rev. tech. Comp. franç. Thomson-Houston*, April 1951, No. 15, pp. 37–42.) A modification of Pound's stabilization circuit (1690 of 1947) is described which avoids the use of magic-T junctions. See also 2758 of 1951.

621.316.761.2 : 621.396.67 **60**

Increase of the Bandwidth of Aerials by means of Compensation Circuits.—Goublin. (See 39.)

621.318.4 **61**

Mean Winding Length.—L. Hubl. (*Elektrotechnik, Berlin*, April 1951, Vol. 5, No. 4, pp. 175–179.) Methods and formulae are given for determining the mean length of choke and transformer windings on laminated cores. Tables and diagrams enable the required values to be found directly for many practical cases without much calculation, the values in the tables being used for interpolation.

621.318.572 **62**

Transition of an Eccles-Jordan Circuit.—J. R. Tillman. (*Radio tech. Dig., Édn franç.*, 1951, Vol. 5, Nos. 3 & 4, pp. 143–146 & 203–216.) French version of 2119 of 1951.

621.318.572 **63**

Electronic Switch.—S. F. Pinasco. (*Rev. telegr., Buenos Aires*, June 1951, Vol. 39, No. 465, pp. 333–335, 353.) Description, with detailed circuit diagram, of a switching circuit consisting of a multivibrator which produces the voltages necessary for alternate blocking and unblocking of the signal, and an amplifier giving the required amplitude.

621.318.572 : 621.385.5 **64**

High-Speed Sampling Techniques.—B. R. Shepard. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 112–115.) Condensed version; for full paper see *Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 255–265. An arrangement is described which is suitable for transmitting observations from a number of measuring elements in rapid cyclic succession, using only one channel; signals of low intensity are considered. A rotating-radial-beam-valve commutator of the type described by Skellett (3167 of 1944) is connected to the individual circuits in turn. The noise generated by this scanning operation is discussed; its magnitude determines the amount of preamplification required. Scanning rates of 10^5 elements/sec are obtainable.

621.319.4 **65**

Working Characteristics and Performance of Ceramic Capacitors.—A. Danzin. (*Ann. Radioélect.*, April 1951, Vol. 6, No. 24, pp. 156–179.) The performance of capacitors of a great variety of types formed from the ceramic dielectrics previously described (132 of 1951) is discussed in detail. Manufacturing processes are described and tables are given analysing the general characteristics: stability, dielectric loss, endurance under varying climatic conditions, temperature coefficient, and overload characteristics.

621.392 **66**

The Effect of Alteration of Network Parameters in a Particular Type of Reactance Network on the Resonance Frequency of the Network.—T. O'Callaghan. (*Frequenz*, Feb. 1951, Vol. 5, No. 2, pp. 44–47.) The networks considered are of low-pass shunt-capacitance ladder type with n unequal capacitors and $n + 1$ unequal inductors. Two cases are distinguished: (*a*) where available component magnitudes do not coincide exactly with design requirements for a desired performance of the network; (*b*) where it is desired to eliminate an undesired resonance occurring in an existing network. The method developed is based on considering half-sections of the network as separate resonant systems.

621.392.001.4 **67**

Determination of Amplitude and Phase Characteristics of Linear Networks by means of Square Waves.—Müller. (See 209.)

621.392.5 : 512.831 **68**

Transformation of the Quadripole-Network Matrix to the Diagonal Form.—H. Schulz. (*Arch. elekt. Übertragung*, June 1951, Vol. 5, No. 6, pp. 257–266.) Description of a method which systematizes all types of linear quadripole, with application to a low-loss transformer with fixed coupling.

621.392.52 **69**

The Four-Circuit Band-Pass Filter and Bandwidth Switching by Back Coupling.—G. Hentschel. (*Funk u. Ton*, June 1951, Vol. 5, No. 6, pp. 281–286.) The formulae for calculating the resonance curves of 4-circuit filters with arbitrary values of coupling and damping are derived. By introducing auxiliary coupling from the fourth to the first circuit, bandwidth switching is effected very simply. The filter than approximates to a 2-circuit filter with enhanced coupling.

621.392.52 **70**

Transfer Properties of Single- and Coupled-Circuit Stages with and without Feedback.—J. W. Muehlner. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 939–945.) "Universal curves of the reciprocal of the complex response function are studied for systems with one or two tuned circuits. The normalized parabola for two circuits is presented with a net of loci for the origin of co-ordinates within it. The ratio Q_1/Q_2, the amount of fixed detuning, the amount of coupling, and amplitude and phase of any additional feedback independently displace the origin with respect to its position in case of two identical, uncoupled circuits without feedback. Complex cases may be easily analyzed. Different methods incorporating regeneration or feedback loops are discussed which allow practical bandwidth control within wide limits."

621.392.52 : 621.395.44 **71**

Group-Combining Filters for Twelve and Twenty-Four Circuit Carrier Systems.—D. W. Robson. (*G.E.C. Telecommun.*, 1948, Vol. 3, No. 1, pp. 45–51.) Complementary high-pass and low-pass crystal filter units are described for combining and separating the two sets of twelve channels in the 12–60-kc/s and 60–108-kc/s bands respectively, in a 24-channel carrier system. An attenuation of > 40 db is achieved for the two stop bands of, respectively, 60·3–108 and 12–59·7 kc/s, with a pass-band loss level to within ± 0·5 db.

621.392.53 **72**

Waveform Systems and Time 'Equalizers'.—D. C. Espley. (*Wireless Engr*, Aug. 1951, Vol. 28, No. 335, pp. 251–258.) A new approach to the solution of circuit problems where the available transmission information is given in terms of distorted waveforms. Some classical transforms relating frequency and time characteristics can be replaced by simpler algebra which enables the physical interpretation to be kept in view. The amplitude and phase characteristics of a system with unknown internal structure are evaluated simply from the non-redundant networks which are arranged to simulate the actual system waveform distortions. The insertion characteristics of the waveform-correcting networks, or 'time equalizers', follow directly. The correction of some systems may require the use of both 'time equalizers' and the better known 'steady-state' equalizers.

621.392.6.062 **73**

N-Terminal Switching Circuits.—E. N. Gilbert. (*Bell Syst. tech. J.*, July 1951, Vol. 30, No. 3, pp. 668–688.) Shannon's method of synthesis of 2-terminal switching networks (2456 of 1949) is generalized to apply to N-terminal networks which use selector switches with any number of positions.

621.396.611.1 **74**

Constant Oscillatory Circuits for High Frequencies.—K. Schreck. (*Fernmeldetech. Z.*, April 1951, Vol. 4, No. 4, pp. 145–149.) The design and properties are described of various fixed and variable circuits including ceramic components, which give a frequency constancy of the same order as that obtained with quartz crystals. The methods of compensation specially developed for new types of circuit are described.

621.396.611.1 **75**

Internal Resonance in Circuits containing Nonlinear Resistance.—R. E. Fontana. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 945–951.) The simultaneous presence of two oscillations locked in synchronism is investigated for circuits containing a nonlinear element. Such a circuit is said to be 'internally resonant', the ratio of the two frequencies being that of two integers. The solution of the appropriate equations and the conditions for stable equilibrium are discussed.

621.396.611.1 **76**

Oscillation Excitation in Circuit with Periodically Varying Capacitance.—W. Taeger. (*Funk u. Ton*, June 1951, Vol. 5, No. 6, pp. 300–309.) Starting from a particular form of the modulation function of the capacitance variation which is rich in harmonics (sawtooth function) and which has the advantage of affording an exact solution of the differential equation describing the phenomena, the conditions are investigated under which the natural damping of the circuit can be eliminated and the currents and voltages follow a simple sine law.

621.396.611.1 **77**

Method of Determining the Forced Oscillations produced by a Periodic E.M.F. of Arbitrary Waveform.—O. M. Bogatyrew. (*Elektrotechnik, Berlin*, May 1951, Vol. 5, No. 5, pp. 194–197.) A formula for the induced current is derived which involves an integral with limits convenient for calculation. The formula is developed by a simple application of the Duhamel integral.

621.396.611.3 **78**

Equations for the Natural Frequencies of Coupled Circuits from the Quadripole Viewpoint.—H. Rukop & H. Kaiser. (*Telefunken Ztg*, June 1951, Vol. 24, No. 91, pp. 64–74.) Various simple and more complex circuits are considered, and formulae for the coupling factors are derived and tabulated together with relations between the circuit parameters. Applications of the theory to circuits involving the various types of coupling are described.

621.396.611.4 **79**

Separation of Degenerate Oscillation Modes in Perturbed Rectangular Cavities.—F. Bosinelli. (*Alta Frequenza*, Oct./Dec. 1950, Vol. 19, Nos. 5/6, pp. 244–258.) A rectangular resonant cavity is considered, the walls of which have a slight arbitrary deformation. The perturbed resonance frequencies and in particular the frequency displacements of two degenerate modes are determined theoretically. The principles followed in the case of twofold degeneracy can be generalized for N degenerate modes. The frequency shifts are calculated for the two degenerate modes TE_{emn} and TM_{emn} in a rectangular cavity which has been deformed into a trapezoidal cavity by tilting one of its walls through a small angle θ. This type of perturbation separates the two modes: to a first approximation the relative frequency shifts are proportional to θ. Similar results are obtained for the two modes TE_{102} and TE_{201} when they are degenerate.

621.396.615 : 621.316.726 **80**

Difference Oscillators and their Constancy.—W. Herzog. (*Arch. elekt. Übertragung*, June 1951, Vol. 5, No. 6, pp. 279–283.) Arrangements are described which produce two frequencies simultaneously and, after rectification, emit the difference frequency. Bridge types of oscillator with similar or dissimilar arms are considered. In the latter case, the use of a crystal as one branch of the circuit results in high constancy of the difference frequency. A simple oscillator is also described which uses a 130-kc/s crystal provided with three electrodes; the resulting difference frequency is 1·3 kc/s.

621.396.615.018.424† **81**

Seven-League Oscillator.—F. B. Anderson. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 881–890.) A bridge-type RC oscillator is described which is continuously variable over a very wide frequency range in one sweep of a two-gang linear-potentiometer control. The output is available in four phases, and the frequency is an approximately logarithmic function of the potentiometer setting. The design of the bridge is discussed in terms of circle diagrams of transmission through RC networks. Tentative frequency limits are 0·01 c/s and 10 Mc/s. Accuracy of setting and stability of frequency are discussed.

621.396.615.029.422/.426 **82**

A Balanced RC Oscillator.—D. A. Bell. (*Electronic Engng*, July 1951, Vol. 23, No. 281, pp. 274–275.) A Wien-bridge feedback circuit with a balanced maintaining amplifier is used to produce frequencies between 0·1 and 20 c/s having a very small even-harmonic content.

621.396.615.14 **83**

Limiting Range for the Generation of Oscillations with Grid-Controlled Valves.—L. Ratheiser. (*Radio Tech., Vienna*, April 1951, Vol. 27, No. 4, pp. 168–170.) Continuation of paper abstracted in 1819 of 1948. Circuit arrangements and construction of various types of u.s.w. oscillator are discussed.

621.396.615.17 **84**

Frequency Multiplier with Selective Network in the Grid Circuit.—A. Taraboletti. (*Alta Frequenza*, Oct./Dec. 1950, Vol. 19, Nos. 5/6, pp. 221–230.) Relative amplitudes of the various harmonics appearing in the anode circuit are calculated for the case of an aperiodic and of a selective grid circuit. Experimental results confirm the superiority of the selective arrangement.

621.396.619.23 **85**

Oscillator Circuits as Frequency Modulators.—Mansfeld. (See 283.)

621.396.645 **86**

Mathematical Treatment of Control Curves for Amplitude-Controlled Amplifiers.—J. Hacks. (*Telefunken Ztg*, March 1951, Vol. 24, No. 90, pp. 51–54.) The dependence of the output voltage on the input voltage is investigated (*a*) for linear dependence on the slope of the control-valve characteristic (hexode with grid-3 control), (*b*) for exponential dependence (pentode, or hexode with grid-1 control). Both cases admit of accurate calculation. Experimental and theoretical curves are compared and the design of a practical i.f. amplifier with forward and backward control is shown.

621.396.645 **87**

Cathode-Follower Input Impedance.—J. E. Flood. (*Wireless Engr*, Aug. 1951, Vol. 28, No. 335, pp. 231–239.) The input conductance G of a cathode follower with a capacitive load decreases with increasing frequency, passes through zero and becomes negative. The circuit with the grid leak returned to a point on the cathode resistance is analysed; an expression is obtained for the frequency at which G is zero. Methods of raising this frequency are discussed; the most satisfactory involves a series grid resistance. Experimental work with a triode-connected EF 91 is described. Discrepancies between predicted and observed behaviour are attributed to Miller effect, resulting from the unavoidable anode load presented by the anode decoupling capacitor.

621.396.645 : 621.392.52 **88**

New Method of Calculating High-Frequency Filters with Tchebycheff Type of Amplification.—H. Edelmann. (*Arch. elekt. Übertragung*, June 1951, Vol. 5, No. 6, pp. 284–292.) With the aid of Tchebycheff polynomials the amount of amplification, and hence the complex amplification, is determined by choice of the null points of the reciprocal of the complex amplification. Circuit parameters can then be chosen so that the null points agree with requirements. The method is applicable to amplifiers with any number of stages and circuits. Typical examples considered are (*a*) single-circuit amplifier stage, (*b*) two-circuit amplifier stage with coupled filters.

621.396.645.018.424† **89**

Design and Construction of Wide-Band I.F. Amplifiers.—L. Gérardin. (*Rev. tech. Comp. franç. Thomson-Houston*, April 1951, No. 15, pp. 19–36.) The general design of i.f. amplifiers for use in radar receivers is discussed. Large bandwidth may be achieved in amplifiers of normal type by the use of certain artifices, but the method has limitations. A distributed amplifier is described which has a pass band of 10–150 Mc/s at ±3 db, with an overall gain of 72 db. The calculation of the overall noise factor of a radar receiver is described in an appendix.

621.396.645.018.78† **90**

Nonlinear Distortion in Push-Pull Class-B Amplifiers with Choke Output.—F. Böttcher. (*Telefunken Ztg*, March 1951, Vol. 24, No. 90, pp. 39–48.) Detailed mathematical analysis of the choke output circuit, with numerical calculations for a 40-kW valve. Approximate formulae are shown to give results which are sufficiently accurate for most purposes.

621.396.645.029.4 : 621.3.018.78† : 534.861 **91**

Audio-Frequency Amplifiers with Adjustable Nonlinearity (Distortion Circuits).—Hoffmann. (See 26.)

621.396.645.2.018.422† **92**

Calculation of High-Frequency Amplifier Circuits with Relatively Narrow Pass Band (Radio Filter Circuits).—W. Cauer. (*Arch. Elektrotech.*, 1951, Vol. 40, No. 2, pp. 88–109.) A general method of calculation is developed whereby any symmetrical frequency characteristic of an amplifier containing only single-circuit coupling networks can be realized in respect of phase and amplitude by an amplifier circuit having a prescribed number of stages with multiple-circuit coupling. Detailed treatment is given for 2- and 3-circuit reactive networks. Application of the method to purely resistive networks is considered.

621.396.645.35 **93**

D.C. Amplifier with Reduced Zero-Offset.—W. McAdam, R. E. Tarpley & A. J. Williams, Jr. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 128–132.) Essentially the same paper as published in *Proc. nat. Electronics Conference Chicago*, 1950, Vol. 6, pp. 277–285. An amplifier for current measurement, using a calibrated 1-MΩ woven-wire resistor, has a zero offset (i.e., input required to bring output to zero) less than 1·6 times the peak-to-peak value of thermal-fluctuation voltages, corresponding to 10^{-12}A on current ranges and 1μV on 2×10^{10} Ω/V voltage ranges. Overall d.c. feedback is used.

621.396.645.37 **94**

Parallel Voltage Feedback.—K. H. R. Weber. (*Elektrotechnik, Berlin*, April 1951, Vol. 5, No. 4, pp. 180–182.) Series and parallel voltage feedback are discussed with the aid of diagrams and are shown to be essentially similar in operation. Parallel voltage feedback involves two damping quantities, which can be determined from formulae given. Amplification with parallel feedback is not directly measurable, but must be calculated from a simple formula involving the two damping quantities and a certain voltage ratio. A particular case is noted in which the output voltage is zero.

621.396.645.371 **95**

Distortion and Power Output of Negative-Feedback Amplifier Output Valves in Class-A Operation.—R. Siegert. (*Telefunken Ztg*, June 1951, Vol. 24, No. 91, pp. 100–117.) A theoretical and experimental investigation, illustrated by many curves, of the distortion remaining after application of negative feedback, first for the case of a phase-constant external resistance and then for a frequency-dependent external resistance such as a loudspeaker.

621.396.645.371 : 621.3.015.3 **96**

Transient Phenomena in Wide-Band Feedback Amplifiers.—O. B. Lur'e. (*Elektrotechnik, Berlin*, April 1951, Vol. 5, No. 4, pp. 171–174.) German version of 78 of 1950.

GENERAL PHYSICS

537.291 **97**

On the Motion of Gaseous Ions in a Strong Electric Field: Part 1.—G. H. Wannier. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 281–289.) The Boltzmann form of the kinetic theory of gases is used to determine the motion of positive ions in a gas under the influence of a static, uniform electric field. The ion density is assumed to be vanishingly low and the field to be of a strength such that the energy it imparts to the ions is not negligible in comparison with their thermal energy. The velocity averages are studied, rather than the details of velocity distribution.

537.52 **98**

Cold Emission of Electrons in Spark Gaps.—F. L. Jones & E. T. de la Perrelle. (*Nature, Lond.*, 28th July 1951, Vol. 168, No. 4265, pp. 160–161.) The results of measurements of the electron emission from a Ni cathode oxidized by previous sparking in air are interpreted as supporting the view that the electrons have their source in the oxide surface layer and are extracted by a field process.

537.52 **99**

Physics of Electrical Discharge.—F. L. Jones. (*Nature, Lond.*, 28th July 1951, Vol. 168, No. 4265, pp. 140–142.) Report of a symposium on 'Some Aspects of Discharge Physics' held on 29th–31st March 1951 at Swansea.

537.52 **100**

The Distribution of Electron Energies in a Discharge constricted by its Self-Magnetic Field.—J. E. Allen. (*Proc. phys. Soc.*, 1st June 1951, Vol. 64, No. 378A, pp. 587–589.) The distribution of electron velocities in a high-current arc in which the radial electric field can be neglected is shown to differ from the Maxwellian, owing to the action of the self-magnetic field. In particular, the proportion of high-velocity electrons is reduced.

537.521 **101**

Microsecond Transient Currents in the Pulsed Townsend Discharge.—J. A. Hornbeck. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 374–379.) The variation of photoelectric emission current with time was observed using an experimental gas-filled tube and a pulsed light source. The method is useful for studying fundamental parameters and processes in the noble gases.

537.525 **102**

Mechanism of Secondary Ionization in Low-Pressure Breakdown in Hydrogen.—F. L. Jones & D. E. Davies. (*Proc. phys. Soc.*, 1st June 1951, Vol. 64, No. 378B, pp. 519–527.) The breakdown mechanism was investigated by examining the dependence of the shape of cathode emission curves on the deposition of electro-positive or electro-negative atoms on cathodes of Ni, Al, Ag, Cu and Mo. "Results are analysed in terms of the varying effective work function of the surface due to the deposition of atoms. It is concluded that the high photoelectric emission from cathodes of low effective work function was due to low-energy photons generated in the pre-breakdown currents, but that for clean metals the secondary emission was due to impact of positive ions; photo-emission due to any high-energy photons generated in the ionization currents was negligible in comparison."

537.525.5 **103**

The Potential Field in and around a Gas Discharge, and its Influence on the Discharge Mechanism.—W. Finkelnburg & S. M. Segal. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 582–585.) Measurements on the cathode and anode fall regions in carbon arcs are described; the thickness of each region is < 0·01 cm. The distorting effect of the nonuniform space-charge distribution on the potential field is discussed. Potential probe measurements in low- and high-current carbon arcs are in good agreement with the theoretical analysis and prove the transitional region between the distorted potential field inside and the undistorted potential field outside the discharge to be a fairly thin one.

537.533 : 538.56 **104**

Radio Wave Generation by Multistream Charge Interaction.—J. Feinstein & H. K. Sen. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 405–412.) Theories of the excitation of plasma oscillations in a two-electron-beam system are reviewed; analytical and graphical methods are used to determine the ranges of parameters within which the growth of waves is possible, taking into account the effect of thermal motion. The growth of waves is possible, within a narrow frequency range, even when the beam injection velocity is much smaller than the mean thermal velocity. Relevance to solar phenomena is discussed. A new theory is advanced of the mechanism by which the longitudinal oscillations are converted into transverse electromagnetic radiations.

537.533 : 538.56 **105**

The Relativistic Theory of Electro-Magneto-Ionic Waves.—V. A. Bailey. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 439–454.) "The relativistic equations governing an ionized gas pervaded by static electric and magnetic fields and the corresponding equations for small perturbations, are derived. The equations for plane perturbations are then obtained and several important cases are developed in detail. Frequency bands in which growing wave-modes occur are also determined. By means of certain rules of transformation the theory is also used to study the plane waves which can occur in interpenetrating double streams of electrons. The results obtained are formally similar to those obtained for waves in an ionized gas. In the absence of static magnetic fields and with the effects of collisions neglected it is found that, in either an ionized gas or in interpenetrating double streams of electrons, certain waves propagated obliquely to the drift motions may both grow and possess Poynting fluxes; these fluxes are such that certain initial disturbances can lead to the escape of amplified electromagnetic energy from an ionized medium. The exchange of momentum and energy, between the streams of electrons and ions and the growing waves, is discussed by means of the momentum-energy tensors of the charged particles and of the electromagnetic field. . . It is concluded that both theory and observation lend support to the hypothesis suggested previously that a notable part of cosmic noise and strong solar noise originates as electro-magneto-ionic waves in magnetized ionized regions."

537.533 : 621.385.029.63/.64 **106**

Waves in Electron Streams and Circuits.—J. R. Pierce. (*Bell Syst. tech. J.*, July 1951, Vol. 30, No. 3, pp. 626–651.) A review of some of the assumptions made and general problems involved in analysis of the properties of electron streams coupled to circuits. The reasons for using a wave approach are explained. "The propagation constant of the wave is obtained in terms of the properties of the electron stream and the impedance of the circuit. Some general properties of waves are described. The importance of fitting boundary conditions in the solution of an actual problem is discussed, and examples, including that of 'backward-gaining' waves," are considered.

537.533.8 **107**

A Theory of Secondary Electron Emission from Metals.—E. M. Baroody. (*Phys. Rev.*, 15th June 1950, Vol. 78, No. 6, pp. 780–787.) A theory is formulated which is based on the Sommerfeld free-electron model, momentum transfer between electrons and lattice being accounted for by introducing a finite mean free path for elastic scattering. An inverse relation between emission and work function is deduced which is in qualitative agreement with experiment. The theory also accounts for the velocity distribution of the secondary electrons and is consistent with their observed angular distribution.

537.533.8 **108**

A Comparison of Theories of Secondary Emission.—J. J. Brophy. (*Phys. Rev.*, 1st June 1951, Vol. 82, No. 5, pp. 757–758.) Comparison of results deduced from the quantum-mechanical theory of Wooldridge (147 of 1940) and the free-electron theory of Baroody (107 above).

537.533.8 **109**

Secondary Emission of Electrons from Liquid Metal Surfaces.—J. J. Brophy. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 534–536.) Curves showing the secondary-emission ratio of Bi, Ga, Pb and Hg as a function of primary energy were obtained experimentally for the metal in both the solid and liquid states; the characteristics for the two states are similar to one another and to the curves for other pure metals. The observed maximum values of secondary-emission ratio are compared with values predicted from theory.

537.533.8 **110**

The Application of Wooldridge's Theory of Secondary Emission.—E. M. Baroody. (*Phys. Rev.*, 15th Aug. 1951, Vol. 83, No. 4, pp. 857–858.) Conclusions are reached in general agreement with those of Brophy (108 above). For Wooldridge's theory see 147 of 1940.

538.123 **111**

The Analytical Expressions for the Vector Potential of a Rotationally Symmetrical Magnetic Field.—A. Moussa & J. Lafoucrière. (*C. R. Acad. Sci., Paris*, 9th July 1951, Vol. 233, No. 2, pp. 139–141.)

538.311 : 621.318.423 : 513.647.1 **112**

The Excitation of a Helical Conductor.—S. Kh. Kogan. (*C. R. Acad. Sci. U.R.S.S.*, 21st Sept. 1950, Vol. 74, No. 3, pp. 489–492. In Russian.) It is assumed that a voltage is applied to an infinitely small element of the helix at L = 0. The distribution of voltage along the helix is then given by equation (2), and the boundary condition, taking into account the finite conductivity of the helix, is given by equation (3). An integral equation (3b) determining the distribution of the current is derived and its solution is reduced to the evaluation of an integral (top of p. 491). For practical purposes it may be assumed that only one wave is propagated along the conductor.

538.566 **113**

Properties of Inhomogeneous Plane Electromagnetic Waves.—G. Bonfiglioli. (*Alta Frequenza*, Oct./Dec. 1950, Vol. 19, Nos. 5/6, pp. 259–266.) Discussion of the propagation of plane-polarized, inhomogeneous waves, i.e., with electric/magnetic field vectors not constant along the wavefront, in an isotropic unbounded space. For any frequency there is an infinite and continuous series of propagation modes. This result is related to the theory of propagation in waveguides and to the theory of total reflection.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.7 **114**

World-Wide Observations of Solar Activity.—(*Tech. Bull. nat. Bur. Stand.*, July 1951, Vol. 35, No. 7, pp. 93–95.) A general account of the National Bureau of Standards' system for collecting information about the various types of solar activity.

523.72 **115**

Measurements of Solar Extreme Ultraviolet and X-Rays from Rockets by means of a $CaSO_4$-Mn Phosphor.—R. Tousey, K. Watanabe & J. D. Purcell. (*Phys. Rev.*, 15th Aug. 1951, Vol. 83, No. 4, pp. 792–797.) Samples of a $CaSO_4$-Mn phosphor, which was insensitive to wavelengths above 1 340 Å, were exposed to sunlight in V2 rockets. Response in the bands 0–8 Å, 1 050–1 340 Å and 1 230–1 340 Å was measured, using Be, LiF and CaF_2 filters. X-rays were observed on one flight which reached 128 km, during which a sudden ionospheric disturbance occurred. Energy at wavelengths between 1 050 and

1 340 Å was observed at heights down to 80–90 km on all four flights. Values of the total radiation intensity in two ranges of the solar spectrum are calculated for a height between 82 and 127 km. The radiation between 795 and 1 050 Å reaching this region has an intensity well above that produced by a 6 000° K black-body sun.

523.72 : 621.396.822 **116**

Solar Radio-Frequency Emission from Localized Regions at Very High Temperatures.—J. H. Piddington & H. C. Minnett. (*Aust. J. sci. Res., Ser. A*, June 1951, Vol. 4, No. 2, pp. 131–157.) The slowly varying component, *S*, of the solar r.f. radiation is correlated with sunspot data and degree of circular polarization. It is suggested that the *S* component is due to thermal emission from localized regions at temperatures of about 10^6°K, often near sunspots. The radiation from a model hot region is examined in detail; the derived emission spectrum and polarization characteristics agree reasonably with observations. Electrons and protons would probably be emitted thermionically from the hot regions, would travel to the earth with average velocities of a few hundred kilometres per second and could form the slow corpuscular radiation deduced from terrestrial magnetic data.

523.72 : 621.396.822 **117**

Solar Noise.—M. Nicolet. (*Scientia*, Feb. & March/April 1950, Vol. 85, Nos. 454 & 455/456, pp. 37–41 & 71–77. In French.) A survey paper; observations on r.f. waves emitted by the sun are described and discussed in relation to known solar and terrestrial phenomena. The noise from the quiet sun originates in the outer regions; centimetre waves originate in the chromosphere and metre waves in the corona. When the sun is disturbed, as indicated by the appearance of sunspots, the noise level rises. With the variation of solar activity bursts are superposed on the steady noise level; these are sometimes extremely intense, being then associated with unusual phenomena such as s.w. fade-outs or eruptions of cosmic rays consisting of charged particles accelerated by the electric field induced by the varying magnetic field of the sunspots. Temperatures of the order of 10^6 deg. K for the quiet sun and 10^{11} deg. K for the most intense bursts are indicated.

523.746 "1950" **118**

Final Relative Sunspot-Numbers for 1950.—M. Waldmeier. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 439–441.)

523.746 "1951.04/.06" **119**

Provisional Sunspot Numbers for April-June 1951.—M. Waldmeier. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, p. 445; *Z. Met.*, Sept. 1951, Vol. 5, No. 9, p. 288.)

523.854 : 621.396.822 **120**

Observations of the Source of Radio-Frequency Radiation in the Constellation of Cygnus.—B. Y. Mills & A. B. Thomas. (*Aust. J. sci. Res., Ser. A*, June 1951, Vol. 4, No. 2, pp. 158–171.) The position and some of the properties of this source were determined. Observed fluctuations in intensity were found to correlate with activity in the ionosphere F region; there is no evidence that the emission from the source varies.

523.854 : 621.396.822 **121**

A Preliminary Survey of the Radio Stars in the Northern Hemisphere.—M. Ryle, F. G. Smith & B. Elsmore. (*Mon. Not. R. astr. Soc.*, 1950, Vol. 110, No. 6, pp. 508–523.) The positions and intensities of 50 discrete sources of radio waves were measured, using an interferometer of high resolving power operating on a wavelength of 3·7 m. The results indicate that most of these radio stars are situated within the galaxy and are distributed with an average density comparable with that of visual stars. They were not identifiable with visual stellar bodies.

538.12 **122**

On the Stability of Magneto-Hydrostatic Fields.—S. Lundquist. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 307–311.) The stability of static magnetic fields in an electrically conducting liquid is investigated mathematically; particular attention is given to the case of twisted cylindrical magnetic fields. Instabilities may be caused by the twisting when it introduces an increase of magnetic energy of the same order of magnitude as the energy of the original field. The discussion is relevant to theories of the earth's magnetic field.

538.711 **123**

Aeromagnetic Survey of Vertical Intensity over the Sound with Apparatus of the BMZ Type.—A. Lundbak. (*Tellus*, May 1951, Vol. 3, No. 2, pp. 69–74.) The magnetometric zero balance was suspended in gimbals in an aircraft flying over the sea area between Denmark and Sweden. Comparison with measurements taken on the ground near the coasts shows that with certain precautions accuracies to within ± 20 gammas can be obtained.

550.38 "1951.01/.03" **124**

International Data on Magnetic Disturbances, First Quarter, 1951.—J. Bartels & J. Veldkamp. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 442–444.)

550.38 "1951.04/.06" **125**

Cheltenham [Maryland] **Three-Hour-Range Indices *K* for April to June, 1951.**—R. R. Bodle. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, p. 445.)

550.384 **126**

Variations of the Magnetic Field and Rotation of the Earth.—N. Stoyko. (*C. R. Acad. Sci., Paris*, 2nd July 1951, Vol. 233, No. 1, pp. 80–82.) From a comparison of recorded data it is concluded that the variation of the geomagnetic field may be explained by variation of the earth's rotation, and that any rotating body should possess a magnetic field of which at least a part is due to the rotation round its axis.

550.385 **127**

Principal Magnetic Storms [April–June 1951].—(*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 446–448.)

551.510.535 **128**

The Modes of Formation of the Ionospheric Layers.—J. H. Piddington. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 409–429.) Ionospheric measurements made during eclipses are analysed and, together with recently acquired solar data, are used in a re-examination of the theories of the ionosphere. It is shown that: (*a*) the rates of disappearance of electrons in all the principal layers may be much higher than hitherto believed; if so, then the number of ionizing quanta needed is also greatly increased; (*b*) there is probably a non-solar source of electron production in region E (and perhaps region D); it may be the electric currents flowing in the upper atmosphere which are also manifested by fluctuations in the earth's magnetic field; (*c*) solar ionizing radiation does not appear to originate (principally) either throughout the corona or near the photosphere, but in regions of rapid temperature transition in the chromosphere and lower corona; excess emission occurs, as has been shown previously, from regions of disturbance, often near sunspots; (*d*) the emission spectrum of these regions may be strong in the quasi-Lyman spectrum of He II and in X rays of wavelength ranging down to about

3Å; during flares, X rays of wavelength less than 1Å may be emitted; (*e*) the F_1 and F_2 ionospheric layers may be accounted for if minor modifications of existing theories are made; the E and D layers may be formed by quanta of a few hundred and a few thousand electron volts, respectively, originating in the chromospheric transition region.

551.510.535 **129**

A Self-Consistent Treatment of the Oxygen Dissociation Region in the Upper Atmosphere.—H. E. Moses & Ta-You Wu. (*Phys. Rev.*, 1st July 1951, Vol. 83, No. 1, pp. 109–121.) A theory is suggested according to which the temperature and the concentrations of oxygen atoms and molecules at a height of 100–140 km can be expressed as functions of the height from a knowledge of the temperature and its gradient at a certain height. The theory is illustrated by an actual calculation; the result obtained differs considerably from those of previous researches.

551.510.535 **130**

Statistical Nature of the Ionosphere.—Ya. L. Al'pert & A. A. Aynberg. (*Zh. eksp. teor. Fiz.*, March 1951, Vol. 21, No. 3, pp. 389–400.) The electromagnetic field of a single signal reflected from the ionosphere is regarded as the resultant of the main reflection and a large number of 'elementary' signals scattered by the ionosphere irregularities. The variations of the latter signals are of a random nature and they characterize the degree of irregularity of the reflecting layer. Formulae are derived for determining the displacement velocity of the irregularities and the ratio of the energy of the main reflection to that of the scattered reflections. Results of experiments are given showing that reflection from the F_2 layer is nearly always of this statistical character.

551.510.535 **131**

Comments concerning the Paper 'Fine Structure of the Lower Ionosphere', by R. A. Helliwell, A. J. Mallinckrodt, and F. W. Kruse, Jr.—R. J. Nertney. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 449–451.) From the results of an extensive series of group-height, absorption, polarization, and phase-height measurements on a frequency of 150 kc/s, a model of the lower ionosphere has been developed which appears to satisfy the theoretical and experimental results on a diurnal and seasonal basis at this frequency. Another interpretation is suggested for the two apparent 'reflection' heights noted by Helliwell et al. (1898 of 1951).

551.510.535 **132**

The D-Layer of the Ionosphere.—A. P. Mitra. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 373–402.) A résumé is given of the present state of knowledge of the D region, both theoretical and experimental work being summarized. New theoretical studies of the formation and structure of the D region are made, based on contemporary knowledge. It is assumed that the D region is produced by ionization of O_2 at the first ionization potential. The densities of electrons and ions are calculated for various heights, taking account of the variation of atmospheric temperature and recombination coefficient with height. The distribution does not follow a Chapman law, the electron density increasing continuously with height, although the ion density has a maximum value at nearly the same height as the rate of ion production, as in the Chapman distribution. The variation of height of the reflecting layer with solar zenith angle for long and very long waves is explained, and satisfactory agreement is obtained between experimental measurements of reflection coefficient at low frequencies and values derived from the theory.

551.510.535 **133**

A Note on Certain Characteristics of the Normal E-Layer.—C. H. Grace. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 452–454.) A method due to Kelso was used in converting E-region equivalent heights, measured at Pennsylvania State College, to true heights. The true height of maximum ionization, h_m, was found not to follow Chapman theory except for an hour after sunrise and an hour before sunset. During most of the day h_m was relatively constant at about 115–120 km. For a period centred on noon, the effective recombination coefficient varies with height, in the range 108–120 km, in a manner suggestive of pressure dependence.

551.510.535 : 523.75 **134**

Ionospheric Effects of Solar Flares.—M. A. Ellison. (*Mon. Not. R. astr. Soc.*, 1950, Vol. 110, No. 6, p. 626.) Summary of paper in *Publications of the Royal Observatory, Edinburgh*, 1950, Vol. 1, No. 4. Five types of sudden ionospheric disturbances, noted during 1949 from continuous radio and magnetic records, were studied in relation to flare outbursts as seen simultaneously by $H\alpha$ light in the spectrohelioscope. The effects on the D-layer are confined to the illuminated hemisphere and are independent of the location of the flare on the sun's disk, showing the operative radiation to be ultraviolet light. The flares may accelerate cosmic-ray particles and these may contribute to the ionization towards the end of the disturbances.

551.510.535 : 621.396.11 **135**

Lateral Deflection of the Ray on Reflection at an Inhomogeneous Ionosphere Layer.—Rawer. (See 244.)

551.510.535 : 621.396.11.029.51 **136**

Effects of Ionosphere Disturbances on Low Frequency Propagation.—Watts & Brown. (See 246.)

551.510.535 : 621.396.11.029.51 **137**

A Method for Obtaining the Wave Solutions of Ionospherically Reflected Long Waves, including all Variables and their Height Variation.—J. J. Gibbons & R. J. Nertney. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 355–371.) A method of solution is given for the one-dimensional wave equation $\pi''/\pi = -\kappa_0^2\epsilon^2(x)$, which arises in the application of wave theory to the study of the ionosphere. The method is applied to the problem of reflection from the E layer at 150 kc/s, Chapman distributions with various maximum electron densities being assumed. A brief comparison with experimental observations indicates that the calculated absorption is too low. This supports the theory of a D region below the E layer. The phase heights are fairly consistent with measured group heights, the sense of the difference agreeing with ionospheric dispersion theory.

LOCATION AND AIDS TO NAVIGATION

621.396.93 **138**

Radio Direction-Finding from the Morphological Viewpoint.—W. Stanner. (*Elektron Wiss. Tech.*, May 1951, Vol. 5, No. 5, pp. 135–176.) A comprehensive review of the subject dealing with (*a*) arrangements for directional reception of medium, long and ultra-short waves, (*b*) arrangements for directional transmission, (*c*) the consol system, (*d*) various methods of distance measurements, (*e*) radar equipment, (*f*) anti-radar methods, (*g*) geophysical and astrophysical problems of radio d.f.

621.396.93 **139**

'Telegon' Direction Finder.—W. Runge, M. Strohhacker & A. Troost. (*Telefunken Ztg*, June 1951, Vol. 24, No. 91, pp. 75–81.) Description of Telefunken equipment for ships, with a wavelength range of 85–1 530 m. Circular crossed coils, 110 cm in diameter, together with a vertical rod 2·6 m long and coincident with the common diameter of the coils, constitute the aerial system. The special features of the Type-E374N receiver are described and a simplified circuit diagram is given. The total frequency range is covered in four ranges with adequate overlap. For an account of the goniometer used see 140 below (Troost & Jankovsky).

621.396.93 **140**

Recent Goniometer Developments.—A. Troost & R. Jankovsky. (*Telefunken Ztg*, June 1951, Vol. 24, No. 91, pp. 81–85.) Improvements in goniometer design resulting from the use of iron-cored rotors, iron rings and shields are reviewed and a description is given of the goniometer used in the Telegon receiver Type E374N (139 above). This is of the iron-shielded type with a high-μ rotor core which gives the requisite inductance of the windings with a relatively small number of turns. The residual bearing error is < 0·15°. Methods of test are outlined.

621.396.93.029.62 **141**

Ultra-Short-Wave Direction-Finding Installation PV-1B (Direction-Finding Installation at Kloten Airport).—W. Schoeberlein. (*Bull. schweiz. elektrotech. Ver.*, 7th April 1951, Vol. 42, No. 7, pp. 226–232. In German.) Outline of the principles of operation and features of the single-receiver automatic system in operation since 1949 [see 1078 of 1949 (Cleaver)]. Sources of error in the Adcock fixed-aerial system are briefly discussed. Remote indication introduces an error of not more than 1%.

621.396.93.088 **142**

Polarization Errors of Radio Direction-Finders, a Proposed Classification.—P. G. Hansel. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, p. 970.) 'Primary instrumental polarization errors' are defined to be those which arise when the response of the aerial system to an 'unwanted' primary component of the received wave is large. 'Secondary polarization errors' are those which arise when the aerial does not respond to the primary 'unwanted' component but takes a bearing on re-radiated waves polarized in the 'wanted' direction.

621.396.93.089.6 **143**

The Calibration of Aircraft Direction-Finders with Particular Reference to Site Selection.—J. H. Moon. (*Marconi Rev.*, 3rd Quarter 1951, Vol. 15, No. 102, pp. 101–112.) Experiments are described which show that radio waves are refracted as they pass over reinforced-concrete runways. Errors were observed on a loop direction finder at frequencies between 200 and 700 kc/s, with waves incident horizontally at an angle of 50° to the line of the runway. The maximum error was about 6° at the edge of the runway and 1·5° at a distance of 100 ft from it.

621.396.933 **144**

A Straight-Line-Flight Indicator for the Pilot of a Radar-Equipped Aircraft.—R. C. Richardson. (*Aust. J. appl. Sci.*, June 1951, Vol. 2, No. 2, pp. 223–234.) An attachment to shoran airborne equipment to assist the pilot in flying along any predetermined straight-line path. Ranges from the ground beacons are transmitted mechanically from the shoran oscillograph to the aircraft equipment, where they are used to set the position of a probe moving parallel to a glass plate coated with a thin layer of metal through which lines are engraved which represent the desired flight paths. Divergence of the flight from one of these lines is indicated to the pilot by a centre-zero meter. In trials the pilot was able to control the aircraft so that its average departure from a straight-line path was about ± 0·02 mile.

621.396.933 **145**

Modern Radar Landing Systems.—M. Sollima. (*HF, Brussels*, 1951, No. 10, pp. 270–286.) The principles and practice of G.C.A. are described and exemplified by the C.F.T.H. installation at Melsbroeck-lez-Bruxelles airport. This uses a precision radar equipment working on a frequency of 9 375 Mc/s with a peak power of 35 kW. The aerials give a horizontal beam 3·5° wide and 0·6° deep for determination of angular elevation, and a vertical beam of width 0·6° and depth 2° for determination of bearing, the energy in secondary lobes being less than 5% of that in the principal lobe. The construction of the variable-width waveguides to secure fine scanning is described, together with ancillary apparatus.

621.396.933 **146**

Blind-Landing System Type A. B.—A. Moisson. (*Rev. tech. Comp. franç. Thomson-Houston*, April 1951, No. 15, pp. 5–18.) A detailed description of the equipment. See also 145 above (Sollima).

621.396.932/.933.24 **147**

Radio Beacon 'Sol'.—M. Borondo. (*Rev. Telecomunicación, Madrid*, Dec. 1950, Vol. 6, No. 22, pp. 23–25.) Short description and theory of the consol system. See also 2912 of 1946 (Clegg) and 2252 of 1948 (Jessell).

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 546.431-31 **148**

Optical Absorption and Photoconductivity in Barium Oxide.—W. W. Tyler & R. L. Sproull. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 548–555.) "Measurements of the optical absorption and photoconductivity in BaO single crystals are reported. The threshold photon energy for both processes is 3·8 eV, and the absorption constant at higher energies is at least 10^5 cm^{-1}. A second increase in the absorption constant begins at about 4·8 eV and is not accompanied by photoconductivity. The observed dependence of photoconductivity on temperature and electric field strength indicates space charge effects in limiting current flow. The onset of absorption at 3·8 eV is thought to be due to the production of excitons, with the accompanying photoconductivity due to thermal dissociation of excitons or exciton ionization of impurity centers."

535.215 : 546.431-31 **149**

Photoconductivity and Photoelectric Emission of Barium Oxide.—H. B. DeVore & J. W. Dewdney. (*Phys. Rev.*, 15th Aug. 1951, Vol. 83, No. 4, pp. 805–811.) Measurements were made on sprayed coatings of BaO on Ni. The variation of the characteristics with temperature and during activation is shown in diagrams. The results are discussed with reference to energy levels.

535.215 + 535.37] : 546.472.21 **150**

A Comparative Study of Photoconductivity and Luminescence.—R. H. Bube. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 393–396.) Measurements are reported of the luminescence and photoconductivity of ZnS as functions of (*a*) operating temperature during excitation, (*b*) time, and (*c*) temperature during thermostimulation. The results are discussed; they indicate different mechanisms for the two processes.

535.215 + 535.37] : 546.482.21 **151**

Electron Mobility and Luminescence Efficiency in Cadmium Sulfide.—L. Gildart & A. W. Ewald. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 359–363.)

535.37 **152**

Electron Penetration and Scattering in Phosphors.—L. R. Koller & E. D. Alden. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 684–685.) ZnS films ranging in thickness from 0·1 to 0·45 μ were deposited on glass and were irradiated at a constant current density of 0·8 μA/cm^2. The luminescence brightness was measured as a function of exciting beam voltage, using a photomultiplier with c.r. tube display. The results show that nearly 90% of the beam energy is lost in a distance of half the range of the electrons.

535.37 **153**

Properties of a CaSO$_4$-Mn Phosphor under Vacuum Ultraviolet Excitation.—K. Watanabe. (*Phys. Rev.*, 15th Aug. 1951, Vol. 83, No. 4, pp. 785–791.) An investigation of the long-period phosphorescence excited by wavelengths below 1 350 Å.

535.37 : 537.39 : 546.281.26 **154**

Injected Light Emission of Silicon-Carbide Crystals.—K. Lehovec, C. A. Accardo & E. Jamgochian. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 603–607.) The yellow light emitted by certain SiC crystals was investigated as a function of temperature and of current through the crystal. The light intensity varies approximately linearly with current, and experiments under pulsed conditions indicate good h.f. response. The emission spectrum changes considerably with temperature but only slightly with current density. The mechanism of the light emission is discussed and attributed to the recombination of carriers injected through *p-n* boundaries. This process is contrasted with emission from phosphors under bombardment.

537.226 **155**

The Origin of Ferroelectricity.—G. A. Smolenski & N. V. Kozhevnikova. (*C. R. Acad. Sci. U.R.S.S.*, 1st Feb. 1951, Vol. 76, No. 4, pp. 519–522. In Russian.) The deciding factor for the appearance of ferroelectricity is a certain mutual disposition of oxygen octahedra in the crystal. A table of all known substances which exhibit this effect and of substances in which this effect can be expected at certain temperatures is included.

537.226 **156**

Domain-Boundary Motion in Ferroelectric Crystals and the Dielectric Constant at High Frequency.—C. Kittel. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, p. 458.)

537.226.1 : 546.431-31 **157**

The Dielectric Constant of Barium Oxide.—R. S. Bever & R. L. Sproull. (*Phys. Rev.*, 15th Aug. 1951, Vol. 83, No. 4, pp. 801–805.) Measurements were made on BaO crystals for the frequency range 60 c/s–60 Mc/s and temperature range −25° to +60°C. The value found for the dielectric constant in these ranges was about 34, rising at the lowest frequencies.

537.311.33 **158**

Volume Rectification in Zincite.—S. R. Khastgir, S. K. R. Tolpadi & S. C. Mitra. (*Nature, Lond.*, 28th July 1951, Vol. 168, No. 4265, pp. 162–163.) Experiments are reported which indicate that the rectification in zincite depends on the volume and not on the area in contact with the electrodes. A theoretical explanation based on the crystal structure is given.

537.311.33 **159**

Semiconductors.—R. L. Ortueta & E. Y. Garrido. (*Rev. Telecomunicación, Madrid*, Dec. 1950, Vol. 6, No. 22, pp. 12–22.) A concise account of the properties of semiconductors, with theory based on energy-level bands. A table and diagrams give the physical constants and rectification characteristics of Ge.

537.311.33 **160**

The Electrical Conductivity of Simple p-Type Semiconductors.—A. Lempicki. (*Proc. phys. Soc.*, 1st June 1951, Vol. 64, No. 378A, pp. 589–590.) An analysis is outlined which establishes a complete analogy between electron and hole conduction. In both cases conductivity passes through a maximum with increasing temperature.

537.311.33 **161**

A Study of Thermoelectric Effects at the Surfaces of Transistor Materials.—J. W. Granville & C. A. Hogarth. (*Proc. phys. Soc.*, 1st June 1951, Vol. 64, No. 378B, pp. 488–494.) The surfaces of various specimens of Ge and PbS have been explored with a whisker contact, and the character of the conduction mechanism in relatively small regions has been examined by the polarity of (*a*) the photovoltaic effect, (*b*) the rectification, and (*c*) the thermoelectric effect. Methods (*a*) and (*b*) give results characteristic of the bulk material as determined by Hall effect measurements, whereas method (*c*) only gives results consistent with (*a*) and (*b*) for uncontaminated cleaved surfaces or for surfaces which have been etched after polishing. For polished surfaces anomalous effects are obtained and point to the existence of a layer on the surface of character different from that of the bulk semiconductor.

537.311.33 : 538.221 **162**

On the Dispersion of Resistivity and Dielectric Constant of Some Semiconductors at Audiofrequencies.—C. G. Koops. (*Phys. Rev.*, 1st July 1951, Vol. 83, No. 1, pp. 121–124.) Semiconducting Ni/Zn ferrites were investigated. A simple model, comprising highly conductive grains separated by layers of lower conductivity, is proposed to explain the dispersion of a.c. resistivity and apparent dielectric constant. Dispersion formulae are given. The theory is supported by experimental results.

537.311.33 : 546.289 **163**

The Effect of Fast Neutron Bombardment on the Electrical Properties of Germanium.—J. W. Cleland, J. H. Crawford, Jr, K. Lark-Horovitz, J. C. Pigg & F. W. Young, Jr. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 312–319.) The effect of lattice disorders introduced by bombardment is discussed from the theoretical standpoint, and measurements are reported. For *n*-type Ge, the conductivity under bombardment first decreases and then increases; on passage through the minimum the material is converted to *p*-type. The conductivity of *p*-type material increases under bombardment, tending towards a saturation value. Figures are quoted for the rate per neutron of carrier removal and introduction in *n*- and *p*-type Ge respectively. Related temperature effects are described. A footnote mentions that for both *n*- and *p*-type Si, bombardment produces only a decrease of conductivity.

537.311.33 : 546.289 **164**

Observations of Zener Current in Germanium *p-n* Junctions.—K. B. McAfee, E. J. Ryder, W. Shockley & M. Sparks. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 650–651.) Junctions in a Ge single crystal were examined using As as the donor impurity and Ga as the acceptor. The reverse I/V characteristic was measured over a 5-decade range of current, and the critical voltage gradient across the junction was studied by observing the variation of capacitance of the junction with reverse voltage. While the slope of the characteristic in the high-current part of the range agrees well with theory, the critical voltage-gradient predicted is about three times the measured value.

537.312.6 : 621.315.592† **165**

Investigation of Semiconductors at High Temperatures. Refractory Thermistors.—N'Guyen Thien-Chi & J. Suchet. (*Ann. Radioélect.*, April 1951, Vol. 6, No. 24, pp. 99–105.) The materials investigated included the normal refractory oxides, MgO, SiO_2, ZrO_2, etc., and mixtures of these with such metals as Ni, Co, Fe, etc., and the semiconducting oxides of Zn, Ti, Ta and V. The work has resulted in the commercial production of a wide range of stable, robust thermistors for use at 250–300°C, possessing high temperature coefficients (5·5%/°C), and of refractory thermistors operating up to 1 100°C, giving much better control than is obtainable with thermocouples or the usual resistive probes.

537.312.8 : 546.289 **166**

The Magnetoresistance Effect in Oriented Single Crystals of Germanium.—G. L. Pearson & H. Suhl. (*Phys. Rev.*, 15th Aug. 1951, Vol. 83, No. 4, pp. 768–776.) An extensive experimental study of the effect as a function of the orientation of magnetic field and electric current relative to the crystal axes. "The measurements are internally consistent with existing phenomenological theory based on cubic crystal symmetry, in which terms involving the magnetic field to higher than the second order are neglected. It is shown that such deviations as do occur arise from higher terms in the field, since an extension of the phenomenological theory to the fourth order predicts their symmetry. Relations are established between the experimentally observed phenomenological constants and those constants appearing in existing magnetoresistance electronic theories."

538.221 **167**

On Ferromagnetic States.—J. Giltay. (*Appl. sci. Res.*, 1951, Vol. B2, No. 3, pp. 199–216.) Reprint. See 126 of 1951.

538.221 **168**

Structure and Properties of Ferrites.—F. G. Brockman. (*Elect. Engng, N.Y.*, June 1951, Vol. 70, No. 6, pp. 489–494.) 1951 A.I.E.E. Winter General Meeting paper. Advances in the understanding of ferromagnetic ferrites subsequent to those reported in 660 of 1950 are reviewed; particular attention is paid to Néel's theory.

538.221 : [534.321.9 : 534.373 **169**

The Influence of Magnetization on Ultrasonic Attenuation in a Single Crystal of Nickel or Iron-Silicon.—S. Levy & R. Truell. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 668–669.) Report of measurements made using ultrasonic beams of frequency 5–50 Mc/s, with constant magnetic fields, of intensities ranging from zero to saturating field, applied both perpendicular and parallel to the beam direction. At saturation the beam attenuation is low and constant over the frequency range.

538.632 : 621.315.592† **170**

Resistivity and Hall Constant of Semiconductors.—C. N. Klahr. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, p. 460.) Correction to paper noted in 2206 of 1951.

539.23 : 546.26 : 537.311.32 **171**

The Conductivity of Thin Carbon Films.—A. Blanc-Lapierre, M. Perrot & N. Nifontoff. (*C. R. Acad. Sci., Paris*, 9th July 1951, Vol. 233, No. 2, pp. 141–143.) Nonlinear films of resistance 1–50 MΩ have been found by Perrot & Lavergne to have different resistance properties over different ranges of values of applied voltage. An experimental investigation of carbon films of resistance $<$ 0·5 MΩ is reported here which extends the range of the earlier findings. Voltages of 1–40 V were used. Results are shown as families of log R/log I curves; formulae for the various ranges are developed and values derived for various parameters are tabulated.

539.23 : 546.59 **172**

Influence of the Support on the Crystallization of Very Thin Gold Films.—A. Colombani & G. Ranc. (*C. R. Acad. Sci., Paris*, 2nd July 1951, Vol. 233, No. 1, pp. 46–48.) Continuation of experimental investigation noted in 2739 of 1951. Support materials studied include KCl, KBr, $NaNO_3$, NaCl, rhodoid, plexiglas, glass.

546.431.824 **173**

The Lorentz Correction in Hexagonal Barium Titanate.—J. R. Tessman. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 677–678.) An explanation based on the structure is advanced for the experimentally observed absence of ferroelectricity in the hexagonal modification of $BaTiO_3$. See also 2188 of 1950 (Slater).

548.52 : 546.431-31 **174**

Growth and Manipulation of Barium Oxide Crystals.—R. L. Sproull, W. C. Dash, W. W. Tyler & A. R. Moore. (*Rev. sci. Instrum.*, June 1951, Vol. 22, No. 6, pp. 410–414.)

549.514.51 : 621.396.611.21.002.2 **175**

Manufacture of Quartz Piezoelectric Crystals.—G. Floréz. (*Rev. Telecomunicación, Madrid*, June 1951, Vol. 6, No. 24, pp. 3–22.) A detailed account of all processes from the initial examination of the raw material and determination of the optic axis up to the deposition of electrode material on the faces of crystal plates and their subsequent mounting.

620.197 **176**

Some Principles and Practices in Tropicalisation.—G. A. Durrance, W. J. D. Harrison & R. T. Lovelock. (*G.E.C. Telecommun.*, 1949, Vol. 4, No. 1, pp. 23–32.) An account of the adverse effects of various climatic conditions on telecommunications apparatus and of the means of overcoming them by impregnation, metal coating, the use of sealed components, etc. The importance of suitable mechanical design is stressed.

621.315.61 **177**

Dielectric Properties of Insulators.—K. W. Wagner. (*Alta Frequenza*, Feb. & April 1951, Vol. 20, Nos. 1 & 2, pp. 3–27 & 68–79.) Theories regarding the relation of the dielectric properties to the structure of the material are reviewed. The influence of temperature and humidity is examined. The effects of drying and impregnation on paper are investigated.

666.1.037 **178**

A Method of Sealing Sapphire to Glass.—H. Rawson. (*J. sci. Instrum.*, July 1951, Vol. 28, No. 7, pp. 208–209.) A modification of the method given in 179 below is described, using h.f. induction heating.

666.1.037 : 621.383 **179**

A Method of Sealing Sapphire to Glass and its Application to Infra-red Photocells.—R. P. Chasmar, J. L. Craston, G. Isaacs & A. S. Young. (*J. sci. Instrum.*, July 1951, Vol. 28, No. 7, pp. 206–207.) The requirements of window materials for PbTe cells are discussed; artificially grown sapphire is both suitable and readily available. In the sealing method described the glass is heated to a high temperature so that it wets the sapphire.

533.5 **180**

La Technique du Vide. [Book Review]—M. Leblanc. Publishers: Armand Colin, 188 pp. (*Bull. Soc. franç. Élect.*, May 1951, Vol. 1, No. 5, p. 252.) ". . . this little treatise will be useful to all those . . . who are interested in vacuum technology."

621.315.616 : 058.2 **181**

Modern Plastics Encyclopedia and Engineer's Handbook, 1951. [Book Notice]—Publishers: Plastics Catalogue Corporation, New York, 15th edn, 636 pp. A record of developments in plastic materials, machinery, engineering, techniques and applications, with a special section, 'Plastics in Defense', giving information on applications of plastics by the U.S. Services and related government bureaux and on the location of the relevant government offices. The directory section lists all U.S. manufacturers of plastic materials, machinery and equipment.

MATHEMATICS

51 : 537.315.6 **182**

Mathematical Tools for the Solution of Problems of the Distribution of Potential.—J. J. R. Moral. (*Rev. Telecomunicación, Madrid,* June 1951, Vol. 6, No. 24, pp. 38–41.) In many cases such problems are most conveniently dealt with in cylindrical or spherical coordinates. Laplace's equation is transformed to spherical coordinates and a solution is obtained in terms of Legendre functions.

512.831 : 621.392.5 **183**

Transformation of the Quadripole-Network Matrix to the Diagonal Form.—Schulz. (See 68.)

681.142 **184**

The New Universal Digital Computing Machine at the University of Manchester.—T. Kilburn. (*Nature, Lond.,* 21st July 1951, Vol. 168, No. 4264, pp. 95–96.) Detailed description of a machine recently put into service, incorporating improvements over the experimental model described in 133 of 1950.

MEASUREMENTS AND TEST GEAR

537.54 : 621.396.822 **185**

Gaseous Discharge Super-High-Frequency Noise Sources.—H. Johnson & K. R. Deremer. (*Proc. Inst. Radio Engrs,* Aug. 1951, Vol. 39, No. 8, pp. 908–914.) The positive column of a discharge in argon provides a standard noise source at microwave frequencies, the noise temperature being about 15·5 db above 290°K. The discharge tube is inserted diagonally across a waveguide, a good termination being obtained without using tuned elements. Performance data are given for the frequency band 2·6–26 kMc/s; the effects of varying the discharge parameters are discussed.

621.3.011.2(083.74) : 621.315.212 **186**

Coaxial Impedance Standards.—R. A. Kempf. (*Bell Syst. tech. J.,* July 1951, Vol. 30, No. 3, pp. 689–705.) "The calibrations of bridge networks used in developmental tests on coaxial cable are obtained by comparison of the networks with calculable standards of impedance consisting of a group of short-length precision copper coaxial lines. The standards are calculable by reason of the availability of precise formulae relating the distributed primary constants to the measurable physical constants and dimensions of the coaxials. This paper outlines the constructional problems and design features of a group of such standards of impedance which provide a range of values over a broad band of frequencies."

621.317.3 : 621.314.6 **187**

Measurement of the Temperature Dependence of the Voltage Drop and of the Backward Current in Dry Rectifiers. Measurement Methods and Apparatus.—K. Maier. (*Arch. tech. Messen,* June 1951, No. 185, pp. T71–T72.) Measurements of the static characteristic are carried out with direct voltage, but most other measurements with pulsed steady direct voltage. A method for the simultaneous measurement of the dynamic forward and backward currents is described.

621.317.3 : [621.396.615.029.3 + 534.232 **188**

The Generation of Audio-Frequency Voltages for Measurement Purposes.—O. Schmid. (*Funk u. Ton,* March 1951, Vol. 5, No. 3, pp. 133–142.) General description of the principles of different types of tone source, including electromechanical devices and particularly self-excited oscillators. Basic circuits are shown for the RC-, glow-discharge-tube- and heterodyne-type generators; their features and advantages are discussed. For high precision, tuning-fork or quartz generators are required.

621.317.3 : 621.396.822 **189**

The Measurement of Fluctuation Noise by Diode and Anode-Bend Voltmeters.—R. E. Burgess. (*Proc. phys. Soc.,* 1st June 1951, Vol. 64, No. 378B, pp. 508–518.) "The response of a diode voltmeter to a c.w. sine-wave signal and to noise is analysed for three typical diode characteristics: discontinuous linear, discontinuous parabolic and exponential. In the first type the indication is proportional to the r.m.s. value of the input voltage but the constant of proportionality is different for noise and for c.w. and depends upon the ratio of the load resistance to diode resistance. For small input signals the curvature of the diode characteristic is important and the voltmeter tends to a square-law behaviour with the result that c.w., noise or a mixture of c.w. and noise can all be measured in terms of a single calibration. The input conductance of a diode voltmeter is in general greater for noise than for c.w. It can be very much greater in practical conditions and this represents a possible source of error in the measurement of noise. The response of the anode-bend voltmeter to c.w. and to noise is considered for a discontinuous parabolic and for an exponential characteristic. The departure of the c.w. calibration from a square law and the difference between c.w. and noise calibrations are evaluated in terms of the characteristics."

621.317.318.087 : 621.395.625.3 **190**

Magnetic-Tape Recording of Electrostatic Field Changes.—N. D. Clarence & D. J. Malan. (*Proc. phys. Soc.,* 1st June 1951, Vol. 64, No. 378B, p. 529.) Brief description of a method of photographic recording which reduces film consumption.

621.317.335.3.087.4† **191**

Automatic Measurement Computation and Recording of Dielectric Constant and Loss Factor against Temperature.—E. B. Baker. (*Rev. sci. Instrum.,* June 1951, Vol. 22, No. 6, pp. 376–383.) The terms ϵ' and ϵ'' are recorded as a function of temperature from − 23°C to + 150°C, at given frequencies between 50 c/s and 600 kc/s, for solids in the form of thin disks. A modified Schering bridge is used; circuit details are given of the associated equipment.

621.317.337.029.64† **192**

Detection of Small Variations in the Quality Factor of a Cavity Resonator.—R. Malvano & M. Panetti. (*Alta Frequenza,* Oct./Dec. 1950, Vol. 19, Nos. 5/6, pp. 231–243.) Description and analysis of a dynamic method in which a f.m. signal is applied to the cavity input. A crystal detector circuit in the output is used to measure the relative variations in output voltage. These are proportional to $\Delta Q/Q$. Errors introduced by the generator and detector are negligible. An example is given of the application of the method in investigation of the absorption spectra of paramagnetic substances.

621.317.35 **193**

Harmonic Analysis of a Distorted Oscillation after Amplitude Limiting.—M. Kolscher. (*Arch. elekt. Übertragung*, June 1951, Vol. 5, No. 6, pp. 293–299.) In f.m. oscillations, distortion can be reduced by amplitude limiting. For the case of a sinusoidal interfering oscillation, this effect has hitherto only been investigated theoretically under particular assumptions concerning the frequency ratio and amplitude ratio of the main and interfering oscillations, and the shape of the limiter characteristic. The dependence of the results on the frequency ratio is here made clear and a new method of analysis is developed which is applicable for all amplitude ratios and characteristics. Numerical calculations for a typical case show what limiter characteristics give the best interference elimination, and what characteristics can be used without the interference becoming appreciably greater.

621.317.42 : 621.383.27† **194**

Investigation of the Conditions for Use of Kubetski's 'Mosaic' Multiplier as an Indicator of the Magnetic Field.—Krasnogorskaya. (See 298.)

621.317.42 : 621.385.833 **195**

Inductance Method for studying the Field Pattern on the Axis of a High-Power Magnetic Electron Lens.—C. Fert & P. Gautier. (*C. R. Acad. Sci., Paris*, 9th July 1951, Vol. 233, No. 2, pp. 148–150.)

621.317.43 **196**

Measurements on Coils with Iron Cores at Frequencies between 1 c/s and 1 Mc/s.—H. Wilde. (*Arch. tech. Messen*, April 1951, No. 183, pp. T39–T40.) Three factors of which account should be taken are hysteresis, magnetic after-effect and eddy currents. Effective means described for determining these are respectively: permeability measurement at very low frequencies using a type of mutual-inductance bridge to eliminate the effect of copper losses; extrapolation to zero field strength of the inductance characteristic measured by a Maxwell-Wien bridge; a similar procedure carried out at high frequency using a capacitance-tuned resonance bridge or one with a folded slide-wire balancing element.

621.317.44 **197**

Principles of the Design of Pulse-Transformer Laminations: Pulse Hysteresis Meter.—P. Bouvier & B. Daugny. (*Rev. tech. Comp. franç. Thomson-Houston*, April 1951, No. 15, pp. 51–67.) A hysteresis meter is described for use in investigating the properties of magnetic cores under operating conditions such as obtain in the pulse transformers used in radar equipment. It consists of a modulating unit of peak power 1 MW producing a 0·5-μs pulse across the terminals of a winding on the core tested, and a measurement unit enabling the shape of the voltage pulse, the magnetic field, the magnetic induction and the hysteresis loop to be obtained.

621.317.7 **198**

Sensitivity of Electrical Measurement Apparatus and its Enhancement.—F. Moeller. (*Arch. tech. Messen*, June 1951, No. 185, p. T68.)

621.317.7 : 537.312.6 : 621.315.592† **199**

The Use of Thermistors for Temperature Compensation in Precision Measurement Apparatus.—N'Guyen Thien-Chi & J. Suchet. (*Ann. Radioélect.*, April 1951, Vol. 6, No. 24, pp. 106–108.) The increase of resistance with temperature of the copper coils of electrical measuring apparatus may be compensated by the use in series, or in parallel, of thermistors possessing negative temperature coefficients.

621.317.7 : 621.396.619.2 **200**

A New Modulation Meter.—H. Müller. (*Telefunken Ztg*, March 1951, Vol. 24, No. 90, pp. 55–60.) A meter is described suitable for monitoring the modulation of broadcasting transmitters and magnetic-tape, disk or sound-film equipment. The amplitude range is − 45 db to + 5 db, with an approximately uniform logarithmic scale, and the frequency range is 40–15 000 c/s. A separate indicating instrument is provided in addition to that in the apparatus. Calibration at 0, 1 and 100% modulation is effected by stabilized voltages derived from the supply unit.

621.317.7.085.34 : 621.383 **201**

Some Points in the Design of Optical Levers and Amplifiers.—R. V. Jones. (*Proc. phys. Soc.*, 1st June 1951, Vol. 64, No. 378B, pp. 469–482.) Sources of instability in optical levers using photoelectric amplifiers are discussed, and the design of a high-precision instrument is described.

621.317.7.088.6 **202**

Compensation of A.C. Instruments for Variations in Frequency.—J. H. Miller. (*Elect. Engng, N.Y.*, June 1951, Vol. 70, No. 6, pp. 494–497.) 1951 A.I.E.E. Winter General Meeting paper. A general solution is presented for compensation over the frequency range up to 5 kc/s; the series resistance of the meter is shunted by a capacitor whose value is rigorously calculated.

621.317.727.088 **203**

Leakage, Stray Capacitance and Inductance in Potentiometer Comparison Measurements with Direct and Alternating Currents.—J. H. Gosselin. (*Rev. gén. Élect.*, June 1951, Vol. 60, No. 6, pp. 244–253.) The effects of insulation faults in d.c. potentiometers are examined; these arise notably in the auxiliary circuit connections, and can be reduced by suitably locating the galvanometer and switch. In a.c. potentiometers, stray capacitance gives rise to leakage currents in quadrature with intercircuit voltages, and inductance effects due to ambient magnetic fields are also significant. Compensation methods are described for reducing relative errors due to these effects to the order of 10^{-5}.

621.317.733 : 621.317.382 **204**

A Self-Balancing Microwave Power Measuring Bridge.—L. A. Rosenthal & J. L. Potter. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 927–931.) 1950 U.R.S.I.–I.R.E. Meeting paper. "The basic techniques of microwave power measurement are discussed, together with principles of self-balancing bridges for that application. A method of measuring the power change by means of a nonlinear 'slave' bridge is presented. Temperature compensation problems are discussed as applied to the meter under consideration. The complete power meter is considered along with operating procedure."

621.317.733.011.21† **205**

Impedance Measurement Bridge for Determination of Aerial Impedances in the Medium-Wave Band.—K. Fischer. (*Elektrotech. u. Maschinenb.*, 15th May 1951, Vol. 68, No. 10, pp. 249–252.) The instrument is designed particularly for measurement of impedance in which the reactive component is large, the resistive component being balanced by capacitive tuning. Particular attention is paid to screening and compensation of stray capacitances: tuning reactors are contained in a separate screened unit. The bridge is fed from a tone-modulated h.f. source enabling an aural-null indication of balance to be used.

621.317.733.011.3/.4 **206**

An Easily Operated Precision Measurement Bridge for

Self-Inductors and Capacitors with Losses.—E. Sorg. (*Funk u. Ton*, April 1951, Vol. 5, No. 4, pp. 202–209.)

621.317.733.025 : 621.385.832 **207**

The C.R. Tube as Indicator for A.C. Bridges.—R. Oetker. (*Frequenz*, Feb. 1951, Vol. 5, No. 2, pp. 33–38.) Conditions are investigated for the use of c.r. tubes as indicators of complex variables in familiar bridge circuits; particular attention is given to sources of error.

621.317.74 : 621.397.5 **208**

Transmission Measuring System.—O. D. Engstrom. (*Bell Lab. Rec.*, June 1951, Vol. 29, No. 6, pp. 264–268.) The apparatus measures phase distortion in the transmission of video signals along transmission lines. The calibrating signal consists of two sine waves with a constant frequency difference of 200 kc/s, the frequency range extending upwards to 3 Mc/s. The phase shift of the resulting 200-kc/s envelope for any frequency pair is compared oscillographically with the phase of the envelope resulting from a standard frequency pair of 892 kc/s and 1 092 kc/s.

621.392.001.4 **209**

Determination of Amplitude and Phase Characteristics of Linear Networks by means of Square Waves.—J. Müller. (*Fernmeldetech. Z.*, May 1951, Vol. 4, No. 5, pp. 211–220.) Detailed theory of the method is given. The construction of a square-wave generator for the range 50 c/s–6 Mc/s is described and also the method adopted for its calibration. The great advantage of the method is that the amplitude and phase characteristics of a network can be determined over a relatively wide band of network frequencies by means of a single square-wave test frequency. Auxiliary equipment comprises a wide-band c.r.o. with timebase circuit for a frequency of 1 Mc/s and deflection proportional to time, and a wide-band amplifier. Examples of the application of the method include (*a*) anode circuit of an amplifier, (*b*) 2-stage low-pass filter, (*c*) l.f. transformer, (*d*) magnetophone, (*e*) band-pass filters. See also 210 below (Meyer-Eppler).

621.392.001.4 **210**

Measurement of the Frequency Characteristics of Linear Systems by Single or Repeated Switching Processes.—W. Meyer-Eppler. (*Fernmeldetech. Z.*, April 1951, Vol. 4, No. 4, pp. 174–182.) Non-stationary methods of measurement are discussed and methods using single pulses, periodic series of pulses, or square waves are described, particular reference being made to low-pass, high-pass and band-pass filters, and resonators.

621.396.621.001.4 **211**

Systematic Analysis of the Properties of Radio Receivers.—Fromy. (See 252.)

621.397.6.001.4 **212**

Traveling-Wave Amplifier Measurements.—F. E. Radcliffe. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 110–111.) Circuits and procedures for transmission and impedance measurements on wide-band amplifiers are described. Frequency sweeps of 500 Mc/s at 4 kMc/s with a repetition rate of 60 c/s are used, and measurements are accurate to within 2%.

621.317 : 621.314.6 **213**

Gleichrichtermesstechnik (Rectifier Measurement Technique). [Book Review]—H. F. Grave. Publishers: Geest & Portig, Leipzig, 1950, 227 pp., 27 DM. (*Elektron Wiss. Tech.*, April 1951, Vol. 5, No. 4, p. IV.) One of a series of monographs which deals with the general theory of different types of rectifier and their application in measurement circuits.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.316.578.1 **214**

A Millisecond Timer.—J. M. Sturtevant. (*Rev. sci. Instrum.*, June 1951, Vol. 22, No. 6, pp. 359–362.) The circuits described produce a time delay of up to 10 sec, accurate to within 1 ms, and can be used as a stop-clock reading to the same accuracy, with manual or electrical control. A test pulse generator for adjustment purposes is also described.

621.365.54/.55† **215**

Fundamentals of High-Frequency Heating.—G. Lang. (*Bull. schweiz. elektrotech. Ver.*, 5th May 1951, Vol. 42, No. 9, pp. 289–303. In German.) After a short explanation of terminology and a historical review, the thermal aspect of the subject is discussed. The principle of induction heating is explained and the importance of the depth of penetration is stressed. Dielectric heating is next considered. Different types of generator are reviewed and their advantages and disadvantages for particular applications are noted.

621.365.54† **216**

Additions to the Generalized Theory of Power Transfer by Induction.—M. van Lancker. (*Bull. Soc. franç. Élect.*, May 1951, Vol. 1, No. 5, pp. 241–251.) Further development of the theory published in 1950 (3112 of 1950), leading to a complete solution of the problem.

621.38.001.8 **217**

A Judgment Box.—W. H. Alexander. (*Electronic Engng*, July 1951, Vol. 23, No. 281, pp. 256–257.) Each factor influencing a decision which must be made is assigned a weight; by summation of voltages corresponding to the ten weighted factors, their weighted mean is shown on a meter.

621.38.001.8 **218**

Electronic Protection for War Plants.—R. Y. Atlee. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 96–101.) Photoelectric, capacitance and acoustic alarm systems are discussed. Illustrations of recent installations, including boundary-protection systems, are given.

621.383.001.8 **219**

Industrial Tri-stimulus Color Matcher.—G. P. Bentley. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 102–103.) Construction and circuit details are given of a sensitive flicker-photometer instrument incorporating a photomultiplier.

621.384.6† **220**

The Yale Linear Electron Accelerator.—H. L. Schultz & W. G. Wadey. (*Rev. sci. Instrum.*, June 1951, Vol. 22, No. 6, pp. 383–388.) Design, construction and performance of the accelerator are described. Independent cavity resonators driven by 500-kW triode power amplifiers are used, and large currents of 10-MeV electrons are produced.

621.384.6† **221**

Operation of a Six-MeV Linear Electron Accelerator.—G. E. Becker & D. A. Caswell. (*Rev. sci. Instrum.*, June 1951, Vol. 22, No. 6, pp. 402–405.) A travelling-wave type accelerator is described, using a single magnetron as power source and operating at a wavelength of 10·5 cm, with a peak power output of 0·9 MW.

621.384.611† **222**

On the Radio-Frequency Requirements of High-Energy Electron Synchrotrons.—T. R. Kaiser. (*Proc. phys. Soc.*, 1st June 1951, Vol. 64, No. 378B, pp. 502–507.)

Taking into account the finite rate of rise, the required r.f. voltage for efficient transition from betatron to synchrotron operation is determined for two machines under construction.

621.384.612.2† **223**

The Synchrocyclotron at Amsterdam: Part 3—The Electromagnet.—F. A. Heyn. (*Philips tech. Rev.*, June 1951, Vol. 12, No. 12, pp. 349–364.) Design, construction and performance of the magnet are described. A method has been found for stabilizing the particle orbits; as a result, 28-MeV deuterons can be produced instead of the planned 25-MeV deuterons. Parts 1 & 2: 2790 of 1951.

621.385.38.001.8 **224**

Measurement of Ignition Delay in Internal-Combustion Engines.—J. Kwasieborski. (*HF, Brussels*, 1951, No. 10, pp. 287–292.) Two thyratrons, fired by impulses derived from the phenomena investigated, control the functioning of a pentode whose anode current is measured.

621.385.38.001.8 **225**

A Simple Photostimulator.—R. G. Bickford & W. T. Moffet. (*Electroencephalography clin. Neurophys.*, May 1951, Vol. 3, No. 2, pp. 251–252.) A pulse generator, using a thyratron, triggers a flash tube, giving 0·5–50 flashes/sec. Construction and component details are given.

621.385.833 **226**

German Society for Electron Microscopy: Third Annual Conference.—(*Nature, Lond.*, 14th July 1951, Vol. 168, No. 4263, pp. 70–71.) Brief report of conference held on 18th–20th May 1951 in Hamburg, at which about 80 papers were presented dealing with the electronoptics and applications of electron microscopes.

621.385.833 **227**

On the Dioptrics of Electrostatic Electron Lenses.—G. Wendt. (*Z. angew. Phys.*, June 1951, Vol. 3, No. 6, pp. 219–225.) By inversion of the function expressing the potential distribution on the axis of the lens, characteristic relations are derived for the electron paths in basic types of accelerating lens and in the three-element independent lens. Substitution of axis potential for axial coordinate in the differential equation for the paraxial rays gives an equation of the Heun type with polynomial coefficients, the solutions of which are Riemann functions of second order with four branch points. For a similar method see *C. R. Acad. Sci., Paris*, 20th Feb. 1950, Vol. 230, No. 8, pp. 734–735 (Régenstreif).

621.385.833 **228**

New Formulae for the Third-Order Aberrations of Magnetic Lenses.—P. Sturrock. (*C. R. Acad. Sci., Paris*, 9th July 1951, Vol. 233, No. 2, pp. 146–147.)

621.385.833 : 621.317.42 **229**

Inductance Method for studying the Field Pattern on the Axis of a High Power Magnetic Electron Lens.—Fert & Gautier. (See 195.)

621.387.4* **230**

A Localizing Geiger Counter.—S. G. F. Frank. (*Phil. Mag.*, June 1951, Vol. 42, No. 329, pp. 612–615.) A method is described for locating ionizing particles, based on the fact that the discharge in a Geiger counter spreads along the wire at a constant speed.

621.387.462† **231**

Behaviour of Space Charge in Diamond Crystal Counters under Illumination: Part 1.—A. G. Chynoweth. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 254–263.)

621.387.462† **232**

Behaviour of Space-Charge-Free Diamond Crystal Counters under Beta-Ray Bombardment: Part 2.—A. G. Chynoweth. (*Phys. Rev.*, 15th July 1951, Vol. 83, No. 2, pp. 264–268.)

621.39.001.11 : 6 **233**

Cybernetics.—J. Loeb. (*HF, Brussels*, 1951, No. 10, pp. 257–269, 286.) The development of the subject as a theory of telecommunication signals is outlined and its application to calculating machines and to servomechanisms is described, together with wider applications in physiology, psychology and political economy.

681.142 **234**

Step Multiplier in Guided Missile Computer.—E. A. Goldberg. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 120–124.) A 4 000-valve analogue computer for simulating the characteristics of a missile-target system includes a high-precision multiplier comprising reversible binary counter and relay-operated conductance networks, for maintaining orthogonality between the reference-axis systems of earth and missile as the computation proceeds.

681.142 **235**

Some Aspects of Electrical Computing.—J. Bell. (*Electronic Engng*, June & July 1951, Vol. 23, Nos. 280 & 281, pp. 213–216 & 264–269.) An account of the application of various devices in the design of equipment for conversion from polar to cartesian coordinates or vice versa, differentiation, integration, multiplication, division, and computation of variable functions relating to ballistics. These devices include (*a*) the ipot (inductive potentiometer), a toroidal coil very uniformly wound on a large ring core and provided with sliding contacts, (*b*) the magslip resolver, a special type of servomechanism.

PROPAGATION OF WAVES

538.566 **236**

'Internal' Reflection in a Stratified Medium with Application to the Troposphere.—G. Eckart & T. Kahan. (*Arch. Elektrotech.*, 1951, Vol. 40, No. 2, pp. 133–140.) German version of 718 of 1951.

538.566 **237**

Properties of Inhomogeneous Plane Electromagnetic Waves.—Bonfiglioli. (See 113.)

538.566 : 535.42 **238**

Diffraction of Electromagnetic Waves near the Earth's Surface in an Optically Inhomogeneous Medium.—R. Schachenmeier. (*Arch. elekt. Übertragung*, June 1951, Vol. 5, No. 6, pp. 267–272.) Making use of the laws of geometrical optics and non-Euclidean geometry, the problem is transformed to render it amenable to calculation by methods already developed. For practical calculations, the actual curvature of the earth's surface is reduced by that of the ray path, when the ordinary laws of diffraction can be used.

621.396.11 + 621.392.2.09 **239**

Characteristic Impedance, Power, Voltage and Current in Transmission along Lines, in Waveguides and in Free Space.—Zinke. (See 31.)

621.396.11 **240**

A V.H.F. Field-Strength Survey on 90 Mc/s.—H. L. Kirke, R. A. Rowden & G. I. Ross. (*Proc. Instn elect. Engrs*, Part III, Sept. 1951, Vol. 98, No. 55, pp. 343–359. Discussion, pp. 378–382.) The effects of the height of the site and of the transmitting aerial and the profile of the transmission path are discussed in relation to field-strength measurements for a transmitter at Wrotham, Kent. In built-up areas the variations are less for horizontal than for vertical polarization. In hilly country the minimum field strength occurs on the near-side slope of a valley and is lower for horizontal than for vertical polarization. Measurements along radials show that, in general, signals of these two polarizations are propagated equally well. By extrapolating from these data, a field-strength contour map of S.E. England has been prepared for a transmitter power of 25 kW and high-gain aerial on a 500-ft mast.

621.396.11 **241**

The Propagation of Metre Radio Waves beyond the Normal Horizon: Part 1—Some Theoretical Considerations, with Particular Reference to Propagation over Land.—J. A. Saxton. (*Proc. Instn elect. Engrs*, Part III, Sept. 1951, Vol. 98, No. 55, pp. 360–369. Discussion, pp. 378–382.) The relative importance of abnormally high refraction near the surface of the earth and of reflection from high-level inversion layers is investigated and illustrated by examples. Of the two mechanisms, the latter is the more likely to give abnormally high field strength at ranges of a few hundred kilometres, especially for low terminal heights. Consideration is also given to the effects of scattering by turbulent eddies in the atmosphere; this is of less importance than other mechanisms of propagation up to distances of 250 km.

621.396.11 **242**

The Propagation of Metre Radio Waves beyond the Normal Horizon: Part 2—Experimental Investigations at Frequencies of 90 and 45 Mc/s.—J. A. Saxton, G. W. Luscombe & G. H. Bazzard. (*Proc. Instn elect. Engrs*, Part III, Sept. 1951, Vol. 98, No. 55, pp. 370–378. Discussion, pp. 378–382.) The statistical distribution of quasi-peak field strength as a function of time was determined for propagation over two paths of respective lengths 110 and 270 km at 90 Mc/s, and over one of length 160 km at 45 Mc/s, for a period of two years. The observed field strengths often considerably exceeded the values corresponding to standard atmospheric refraction; some degree of correlation was found with meteorological data obtained from routine radiosonde ascents. Use of the results in planning v.h.f. broadcasting services is outlined. Part 1: 241 above.

621.396.11 : 551.510.535 **243**

Vertical Propagation of Electromagnetic Waves in the Ionosphere.—M. N. Saha, B. K. Banerjea & U. C. Guha. (*Proc. nat. Inst. Sci., India*, May/June 1951, Vol. 17, No. 3, pp. 205–226.) A discussion of the equations for vertical propagation in the ionosphere is given in standardized notation. The electric vector components E_x and E_y are coupled by polarization terms ρ_1 and ρ_2 which are functions of geomagnetic latitude and height; and the propagation vectors associated respectively with the ordinary and extraordinary waves are given by $(E_x + i\rho E_y)/\sqrt{1 + \rho^2}$ for two different values of ρ, and are governed by two refractive indices q_o and q_e and a coupling term $\dot{\phi}$. The five quantities needed to define wave propagation completely are ρ_1, ρ_2, $\dot{\phi}$, q_o and q_e; the first three of these are discussed here, the last two have been dealt with by Booker (see e.g. 355 of 1935) and others. For F-layer propagation $\dot{\phi}$ can be neglected everywhere except very near the geomagnetic poles; E-layer propagation is more difficult to calculate.

621.396.11 : 551.510.535 **244**

Lateral Deflection of the Ray on Reflection at an Inhomogeneous Ionosphere Layer.—K. Rawer. (*Z. angew. Phys.*, June 1951, Vol. 3, No. 6, pp. 226–227.) General and numerical calculation of azimuth variation caused by lateral ray displacement, based on particular ionosphere models. The deflection is in the direction of stronger ionization, and angular displacement of nearly 10° occurs, corresponding to an increase of 10% in the m.u.f. Good agreement with observations is noted.

621.396.11.012.3 **245**

Field Power Conversion.—R. E. Perry. (*Electronics*, Aug. 1951, Vol. 24, No. 8, p. 134.) A chart is shown for converting propagation data from field-strength to power-density values.

621.396.11.029.51 : 551.510.535 **246**

Effects of Ionosphere Disturbances on Low-Frequency Propagation.—J. M. Watts & J. N. Brown. (*J. geophys. Res.*, Sept. 1951, Vol. 56, No. 3, pp. 403–408.) Some photographic records of ionospheric reflections obtained with a transmitter giving a peak pulse power of 900 kW on frequencies of 50, 100 and 160 kc/s are reproduced and discussed. The disturbed appearance of the E-region night-time echoes is found to be a sensitive indication of storminess. In the daytime the effect of storms is to cause an increase in absorption, with a corresponding weakening or fade-out of E echoes. A graph is given showing correlation between magnetic activity and night-time disturbance of the E layer.

621.396.11.029.51 : 551.510.535 **247**

A Method for Obtaining the Wave Solutions of Ionospherically Reflected Long Waves, including all Variables and their Height Variation.—Gibbons & Nertney. (See 137.)

RECEPTION

519.272.15 : 621.39.001.11 **248**

The Electronic Correlator.—H. Doizelet. (*Radio tech. Dig., Édn franç.*, 1951, Vol. 5, No. 4, pp. 187–200.) The problem of improving amplifier sensitivity in spite of irreducible background noise is considered in relation to statistical information theory. The autocorrelation function for noise voltage vanishes for large values of the time interval, while the autocorrelation function for a periodic voltage is itself periodic; this affords a method for separating the two voltages. The operation of the correlator constructed at the Massachusetts Institute of Technology is described. See also 730 of 1951 (Lee, Cheatham & Wiesner).

621.396.61/.62 : 621.396.67 **249**

Transmission and Reception of Circularly Polarized Microwaves with a Common Aerial.—Ruppel. (See 281.)

621.396.62 + 621.395.625] : 061.4 **250**

The Radio Exhibition at the Vienna Spring Fair.—(*Radio Tech., Vienna*, April 1951, Vol. 27, No. 4, pp. 195–198.) Review of portable receivers and magnetophone recording equipment exhibited.

621.396.621 **251**

Short-Wave Receivers.—H. Flicker & J. Hacks. (*Telefunken Ztg*, March 1951, Vol. 24, No. 90, pp. 27–38.) An account of present-day technique used in the construction of commercial s.w. receivers. The properties that can reasonably be expected in a receiver are considered and also how far such requirements are mutually exclusive or concordant. Technical data for twelve modern receivers of German, French, British and American manufacture are tabulated. The propagation of short waves is discussed in relation to receiver characteristics.

621.396.621.001.4 **252**

Systematic Analysis of the Properties of Radio Receivers.—E. Fromy. (*Onde élect.*, May & June 1951, Vol. 31, Nos. 290 & 291, pp. 210–222 & 282–291.) A technique of receiver performance evaluation is described which overcomes the inadequacy of the usual specifications of sensitivity. For the receiver h.f. amplifier a c.w. source is used to derive a family of curves, relating aerial input voltage to detector output, for gradually increasing values of amplification; this is linked with a corresponding family of curves for set noise as a function of input voltage. The a.f. amplifier, with gain control at maximum and the controls in earlier stages adjusted so that set noise at the output terminals is inappreciable, is treated similarly, its characteristics being obtained by varying the voltage applied to the receiver input terminals. The interpretation of the performance characteristics so obtained is discussed and examples are given.

621.396.621.54 : 621.3.012.3 **253**

New Diagrams for 'Ganging' Calculation.—Kerbel. (See 53.)

621.396.622 + 621.396.619.2 **254**

Modulators, Frequency Changers and Detectors using Rectifiers with Frequency-Dependent Characteristics.—Tucker. (See 282.)

621.396.622 **255**

The Relative Advantages of Coherent and Incoherent Detectors: A Study of their Output Noise Spectra under Various Conditions.—R. A. Smith. (*Proc. Instn elect. Engrs*, Part III, Sept. 1951, Vol. 98, No. 55, pp. 401–406.) Summary of I.E.E. Monograph No. 6. Comparison is made of output spectra obtained with incoherent square-law and linear detectors on the one hand and coherent homodyne and commutator detectors on the other hand, for an input comprising sinusoidal signal plus noise. Input noise from several common types of filter is considered. Curves give the signal/noise ratio at the output when no further bandwidth limitation is applied and when the detector is followed by a narrow-band filter. Considerable advantage can be obtained with a coherent detector when the input signal/noise ratio is somewhat less than unity.

621.396.8.029.45 **256**

Overseas Reception on Medium Waves.—K. M. Schwarz. (*Radio Tech.*, *Vienna*, May 1951, Vol. 27, No. 5, p. 224.) Near-East stations often heard in Europe are listed. Reception of medium-wave programmes from S. America is subject to long-period fading (up to several minutes), but is considerably more regular than for s.w. stations.

621.396.823 **257**

Radio Influence Tests in Field and Laboratory — 500-kV Test Project.—G. D. Lippert, W. E. Pakala, S. C. Bartlett & C. D. Fahrnkopf. (*Elect. Engng, N.Y.*, June 1951, Vol. 70, No. 6, pp. 481–486.) 1951 A.I.E.E. Winter General Meeting paper. Report of an investigation of the effect on the unwanted r.f. radiation from 500-kV lines of weathering, precipitation and variation of voltage, frequency and conductor diameter.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 **258**

A History of the Theory of Information.—E. C. Cherry. (*Proc. Instn elect. Engrs*, Part III, Sept. 1951, Vol. 98, No. 55, pp. 383–393.) Fundamental points in human communication systems which have been summarized by precise mathematical theory are mentioned. The need for economy, which appeared as telegraphy and telephony developed, led to systems of signal compression and early theories of communication, later extended to express 'information' quantitatively. Definite accomplishments in the mechanization of processes analogous to thought, as embodied in calculating machines and servomechanisms, are outlined. Evidence from the whole field of scientific observation supports the view that "information plus entropy is an important invariant of a physical system."

621.395.44 **259**

German Carrier-Frequency System V60.—W. Zerbel. (*Fernmeldetech. Z.*, May & June 1951, Vol. 4, Nos. 5 & 6, pp. 193–201 & 268–274.) The carrier-frequency arrangements which will be used by the German Post Office on its new long-distance cable network are described, with particular reference to the electrical aspects (pre-grouping or pre-modulation system for 60 channels) and the practical construction.

621.395.665.1 **260**

Instantaneous Compandors.—C. O. Mallinckrodt. (*Bell Syst. tech. J.*, July 1951, Vol. 30, No. 3, pp. 706–720.) Discussion of the theory of the 'instantaneous' type of compandor, and evaluation of the noise advantage when instantaneous companding is applied to telephone channels.

621.396.5 : 621.396.931 **261**

Experimental Radio-Telephone Service for Train Passengers.—N. Monk. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 873–881.) A series of stationary transmitters and receivers working in the 35–45-Mc/s band and situated along the 440-mile railway from New York to Buffalo provide links between trains and the normal telephone system. Train attendants were formerly required, but experiments with coin-boxes have been successful; in this case the frequency band 152–162 Mc/s is used.

621.396.61/.621 : 621.396.931 **262**

New Equipment for V.H.F. Radio-Telephone Systems.—(*G.E.C. Telecommun.*, 1948, Vol. 3, No. 1, pp. 5–14.) A.m. and f.m. transmitter-receivers for mobile and central stations are described. All units are crystal controlled and will operate at any frequency in the bands allocated between 30 and 184 Mc/s. Miniature valves and components are used, enabling the complete equipment for one station to be contained in a box 8 in. × 18 in. × 8 in. Public-address facilities are incorporated.

621.396.61/.621 : 621.396.931 **263**

High-Frequency Radio Transportable Transmitter and Receiver.—(*G.E.C. Telecommun.*, 1949, Vol. 4, No. 1, pp. 12–22.) A description of apparatus suitable for c.w. or m.c.w. telegraphy and a.m. telephony; the weight is 60 lb. Power supply may be at 12 or 24 V d.c., or from 50-c/s mains. The transmitter, covering the frequency range 2–9·1 Mc/s, comprises a crystal oscillator and modulated amplifier stages and radiates 25 W on c.w. telegraphy. The modulation amplifier may be used alternatively for public-address purposes. The superheterodyne receiver covers the band 2–20 Mc/s.

621.396.619.14 **264**

Product Phase Modulation and Demodulation.—D. B. Harris. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, p. 907.) Correction to paper noted in 3159 of 1950.

621.396.65 **265**

New Directional [radio] **Links.**—J. Kornfeld. (*Radio Tech.*, *Vienna*, April 1951, Vol. 27, No. 4, pp. 164–167.) Review of American and European wide-band relay networks for long-distance communications and television services.

621.396.65 **266**

Application of Microwave Channels.—R. C. Cheek. (*Elect. Engng, N.Y.*, June 1951, Vol. 70, No. 6, pp. 500–503.) 1951 A.I.E.E. Winter General Meeting paper. Relevant factors in the design of microwave communication systems are considered, including the selection of frequency band, the influence of topography on the choice of terminal sites, the calculation of the free-space loss between aerials, which determines the aerial size, and the losses in cables or waveguides between the equipment and aerials.

621.396.65 : 621.396.619.13 **267**

Asymmetry of the Frequency Swing, particularly in the case of Wide-Band F.M. Directional Links.—P. Barkow. (*Fernmeldetech. Z.*, April 1951, Vol. 4, No. 4, pp. 168–173.) The cause of the asymmetry and of the consequent distortion is analysed and methods of correcting and of compensating the errors of symmetry are described.

621.396.65.029.62 : 621.396.5 **268**

Conditions for the Development in France of Two-Way Radiotelephony on Very Short Waves.—E. P. Courtillot. (*Onde élect.*, May 1951, Vol. 31, No. 290, pp. 223–232.) The use of the frequency band 25–174 Mc/s for two-way communication between fixed and mobile stations is discussed generally in the light of current practice in the United States and in England. It is essential to arrange development according to a fixed plan, taking full account of economic factors.

621.396.97 : 621.396.66 **269**

The Automatic Monitoring of Broadcast Programmes.—H. B. Rantzen, F. A. Peachey & C. Gunn-Russell. (*Proc. Instn elect. Engrs*, Part III, Sept. 1951, Vol. 98, No. 55, pp. 329–340. Discussion, pp. 340–342.) See 1491 of 1951.

SUBSIDIARY APPARATUS

621.314.1.082.72 **270**

Electrostatic Direct-Current Transformer of 300 Kilovolts.—J. M. Malpica. (*Rev. sci. Instrum.*, June 1951, Vol. 22, No. 6, pp. 364–369.) A continuous transfer of electric charges is effected by rotating a dielectric between two pairs of metallic brushes, the primary pair being connected to a h.v. source. Several disks on one shaft are used to form a step-up transformer. The elementary theory and experimental results are given.

621.314.632.1 **271**

Resistance Stratification in Copper-Oxide Rectifiers: Part 1—Investigations on Massive Plates with Nonblocking Contacts.—F. Rose. (*Ann. Phys., Lpz.*, 30th June 1951, Vol. 9, Nos. 2/4, pp. 97–123.) Measurements of the apparent resistance of massive oxidized copper disks with graphite or silver contacts were carried out at frequencies up to 100 kc/s and at various temperatures in the range 195°–303°K. The results obtained are presented in a table and diagrams, and are discussed with particular reference to Lehovec's method of estimating the conductivity distribution through the oxide layer.

621.314.632.1 **272**

Resistance Stratification in Copper-Oxide Rectifiers: Part 2—Investigations on Rectifier Disks.—F. Rose. (*Ann. Phys., Lpz.*, 30th June 1951, Vol. 9, Nos. 2/4, pp. 124–140.) Measurements were made for the same frequency and temperature ranges as for massive disks (271 above) on rectifier disks of two sorts of copper subjected to various annealing conditions, again using graphite or silver electrodes. The results are shown in diagrams and Lehovec's method of analysis is applied.

621.316.722.078.3 **273**

Stabilization of Direct Voltages.—H. Günther. (*Funk u. Ton*, March 1951, Vol. 5, No. 3, pp. 124–132.) Discussion of the maximum stability attainable using an amplifier valve as a series resistance. Suitable conductance characteristics for the controlling pentode are shown; its anode load resistance should be as large as possible. An additional potentiometer-type control circuit at the input reduces input voltage fluctuations and reduces the effective internal resistance of the control circuit.

621.319.3 **274**

New Electrostatic Generators.—N. J. Félici. (*Onde élect.*, May 1951, Vol. 31, No. 290, pp. 205–209.) The general principles underlying the functioning of rotary e.s. generators are outlined. The necessity for a fluid dielectric with high breakdown voltage has led to the use of compressed gas, usually air; pressures of 30 atmospheres enable the power output of a machine to be made 200 times greater than is possible with normal atmospheric pressure. Figures are quoted for several existing machines, giving outputs of up to 100 μA at 500 kV. Machines are being developed for outputs as high as 30 mA at 500 kV or more. An overall efficiency of 75% is readily obtained. See also 999 of 1951 (Hémardinquer).

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 **275**

The Evaluation of Picture Quality with Special Reference to Television Systems: Part 1.—L. C. Jesty & N. R. Phelp. (*Marconi Rev.*, 3rd Quarter 1951, Vol. 15, No. 102, pp. 113–136.) "A new method of assessing the performance of a picture-reproducing system is described. The effect of the simultaneous variation of the four parameters, brightness, contrast, resolution, and viewing distance, has been explored. Measurements have been made under conditions of best picture reproduction, with the system maintained at this level of adaptation. Various photographic and television systems have been examined. This work has involved a similar investigation of the behaviour of the 'average observer'."

621.397.5 : 535.62 **276**

A Color Television System for Industry.—H. R. Smith, A. L. Olson & R. F. Cotellessa. (*Elect. Engng, N.Y.*, June 1951, Vol. 70, No. 6, p. 517.) Summary of 1951 A.I.E.E. Winter General Meeting paper. Description of a wired system using an orthicon camera tube and suitable for industrial, commercial, medical and military applications. A field-sequential system with rotating colour disks is used.

621.397.5 : 535.62](083.74) **277**

Plans for Compatible Color Television.—D. G. F. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 90–93.) System standards recommended by the U.S. National Television System Committee are outlined. A composite signal is suggested, including a monochrome component and a colour component on a subcarrier; field tests are to be made in order to formulate definite numerical proposals for a full set of compatible colour standards.

621.397.5 : 535.623 **278**

Dot-Interlaced Scanning and its Recent Development in American Colour Television.—E. Schwartz. (*Fernmeldetech. Z.*, June 1951, Vol. 4, No. 6, pp. 243–250.) The 4-raster system of point scanning is explained and new pulse technique for this method of scanning is described. Its application to colour television is examined in detail and suitable tubes developed in U.S.A. for the purpose are briefly described.

621.397.5 : 621.317.74 **279**

Transmission Measuring System.—Engstrom. (See 208.)

TRANSMISSION

621.316.726.078.3 **280**

U.H.F. Discriminator and its Application to Frequency Stabilization [of klystron].—Pircher. (See 59.)

621.396.61/.62 : 621.396.67 **281**

Transmission and Reception of Circularly Polarized Microwaves with a Common Aerial.—W. Ruppel. (*Fernmeldetech. Z.*, June 1951, Vol. 4, No. 6, pp. 251–253.) By using both directions of rotation of circularly polarized waves an aerial system can be used for both the transmission and reception of waves of the same frequency. The bridge arrangement for decoupling the energy paths is explained and a practical example is described which embodies two tank circuits. Arrangements of compact linear elements avoid the use of a gas-filled blocking valve. The principle is in practice only applicable to microwaves.

621.396.619.2 + 621.396.622 **282**

Modulators, Frequency Changers and Detectors using Rectifiers with Frequency-Dependent Characteristics.—D. G. Tucker. (*Proc. Instn elect. Engrs*, Part III, Sept. 1951, Vol. 98, No. 55, pp. 394–398.) An approximate method of allowing for frequency dependence of the resistance characteristic of the rectifier can be applied easily to the analysis of circuits where the terminating impedances are finite only at a finite number of modulation-product frequencies and zero at all other frequencies. Results for a particular set of operating conditions are illustrated by numerical examples based on copper-oxide modulators working at frequencies up to 6 Mc/s. The conversion insertion-loss can be stabilized against temperature variation by a suitable choice of terminating resistance. The working and results are the same for shunt, series and ring modulators for conditions of low input frequency with high output frequency and vice versa.

621.396.619.23 **283**

Oscillator Circuits as Frequency Modulators.—W. Mansfeld. (*Funk u. Ton*, June–Aug. 1951, Vol. 5, Nos. 6–8, pp. 309–316, 351–360 & 411–421.) Three different arrangements are investigated theoretically and experimentally and it is shown that proportional frequency variations can be obtained by modulation of the feedback phase angle. Using a single-valve oscillator circuit a frequency swing of over 6% was obtained, with linearity over a wide range, but a disadvantage of this modulator circuit is the dependence of the excited frequency on the operating voltages and valve characteristics. A push-pull circuit with push-pull modulation gave much better results, the linearity, frequency swing obtainable, small amplitude variation with modulation voltage, and frequency constancy with variations of operating voltage, making such an arrangement useful as a frequency modulator. Another push-pull circuit, incorporating a phase-control arrangement, also gave good results.

VALVES AND THERMIONICS

621.383.27† : 621.317.42 **284**

Investigation of the Conditions for Use of Kubetski's 'Mosaic' Multiplier as an Indicator of the Magnetic Field.—N. V. Krasnogorskaya. (*Bull. Acad. Sci. U.R.S.S., sér. géogr. géophys.*, Jan./Feb. 1951, Vol. 15, No. 1, pp. 43–50. In Russian.) A multiplier with a CuS-Cs photocathode and six stages of electron multiplication is described. Each stage consists of a secondary-emission CuS-Cs electrode and a nonemitting Ag electrode. The alternation of dark emitting and bright nonemitting layers resembles a mosaic. The electrons leaving the photocathode are directed by means of an electric and a magnetic field; an additional magnetic field shifts the beam from sensitive to non-sensitive electrodes, thus altering the value of the anode current. Experiments with these multipliers are described. The necessary conditions of stability of the power supplies are derived for a given accuracy of measurement of the intensity of the magnetic field, and limits of possible measurements are established.

621.385.029.63/.64 : 537.533 **285**

Waves in Electron Streams and Circuits.—Pierce. (See 106.)

621.385.3 : 546.289 **286**

Electric Forming in *n*-Germanium Transistors using Phosphorus-Alloy Contacts.—J. P. Stelmak. (*Phys. Rev.*, 1st July 1951, Vol. 83, No. 1, p. 165.) Results of measurements emphasize the important role played by the P content of the collector contact point in improving transistor gain when forming is done by pulsed currents.

621.385.5 : 621.318.572 **287**

High-Speed Sampling Techniques.—Shepard. (See 64.)

621.385.82.029.3 : 621.395.61 **288**

Increasing the Efficiency of the High-Power Thermionic Cell by Superposition of a Strong Field obtained from a High Voltage of High Frequency.—S. Klein. (*C. R. Acad. Sci., Paris*, 9th July 1951, Vol. 233, No. 2, pp. 143–145.) Modifications to the gas-filled thermionic cell described in 593 of 1947 are proposed. A h.f. voltage applied to a Pt point produces a discharge which heats the ion emitter. The device can be used both as an acoustic and ultrasonic transducer and as a source of ultraviolet radiation.

621.396.615.141.2 : 621.385.029.63/.64 **289**

The Magnetron as a Travelling-Wave Valve.—O. Döhler. (*Funk u. Ton*, March & May 1951, Vol. 5, Nos. 3 & 5, pp. 146–155 & 257–262.) Treatment of the theory, and description of the construction and operation, of different types of magnetron; complementary to paper noted in 1017 of 1951. See also 2064 of 1950 (Warnecke et al.).

621.38 **290**

Electronics. [Book Review]—P. Parker. Publishers: Arnold, London, 1050 pp., 50s. (*J. sci. Instrum.*, July 1951, Vol. 28, No. 7, p. 223.) "In this admirably balanced work there is no subject which is over-elaborated, or in which the reader is not brought up to a level just short of research standard."

MISCELLANEOUS

621.39.001.5 **291**

Telecommunications Research: Fundamental Investigations undertaken by the D.S.I.R.—(*Wireless World*, Oct. 1951, Vol. 57, No. 10, pp. 431–432.)

621.396 : 061.4 **292**

Radio Exhibition Review.—(*Wireless World*, Oct. 1951, Vol. 57, No. 10, pp. 384–395.) Trends in the design of radio, television, and associated apparatus shown at the 18th National Radio Exhibition, held in London during September 1951, are illustrated and discussed.

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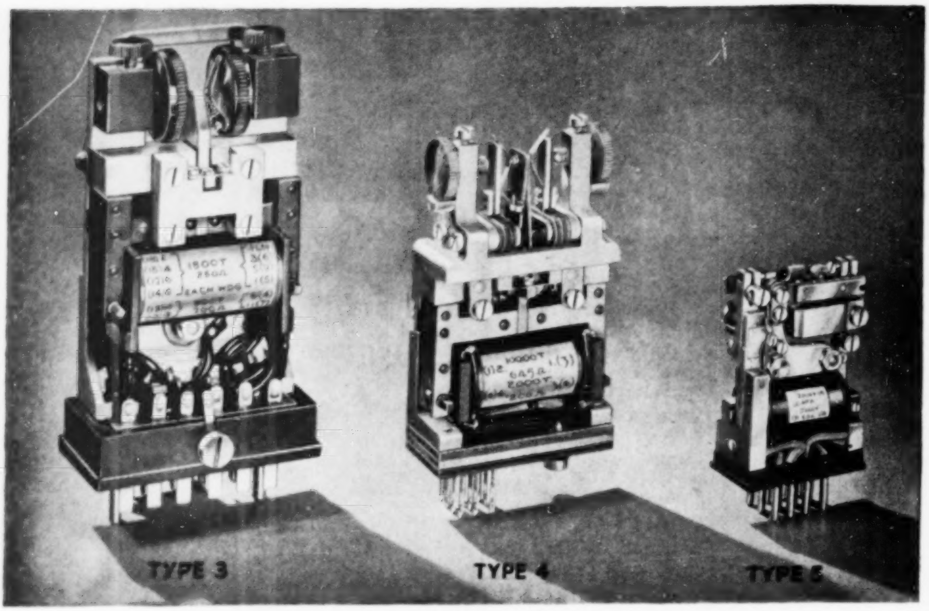

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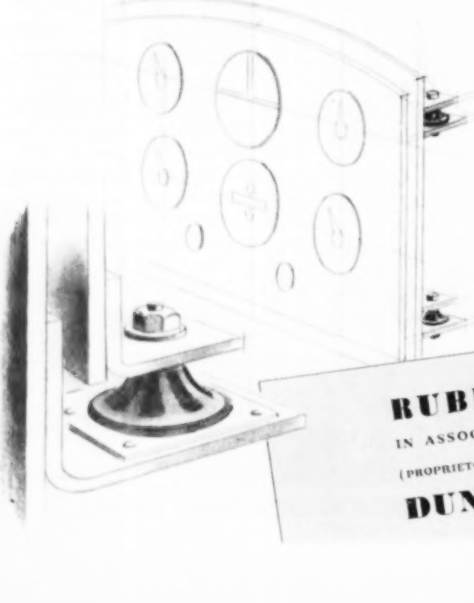

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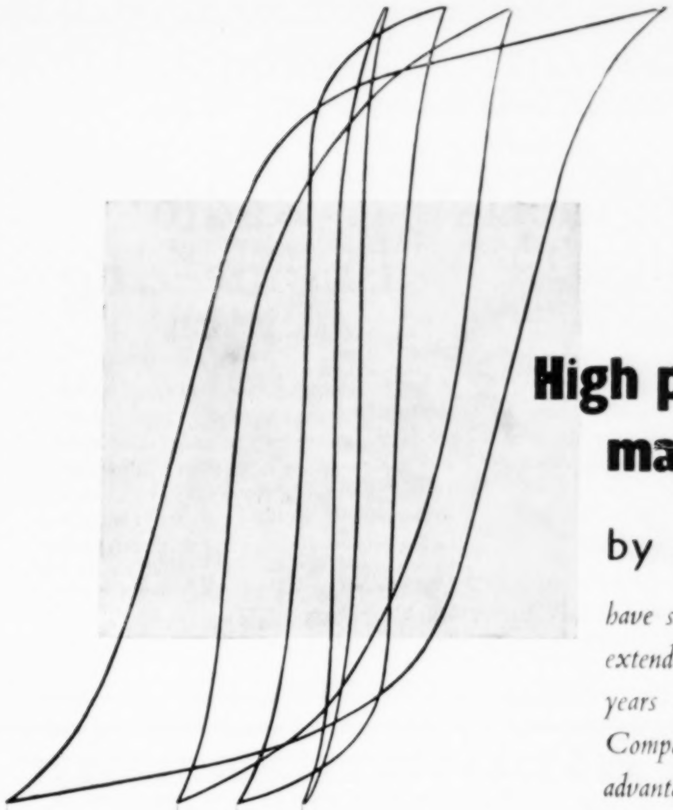

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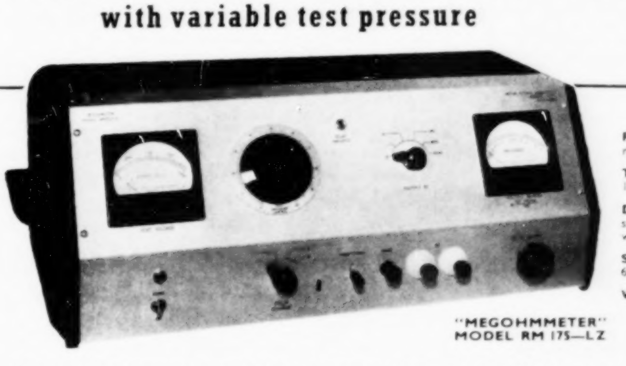

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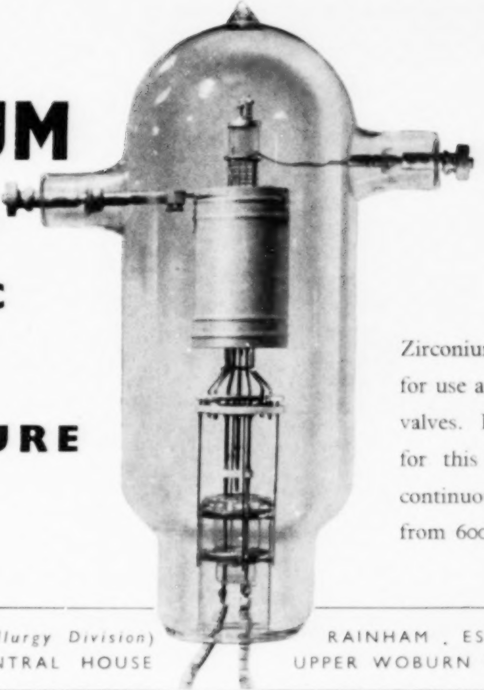

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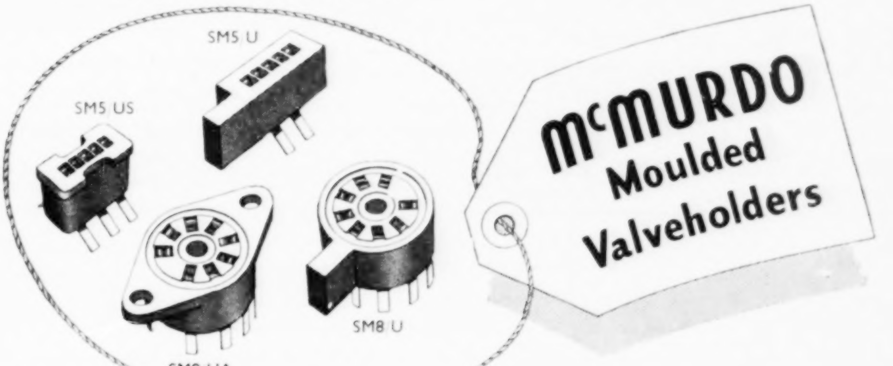

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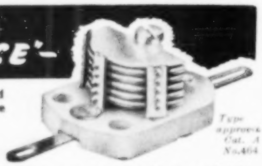

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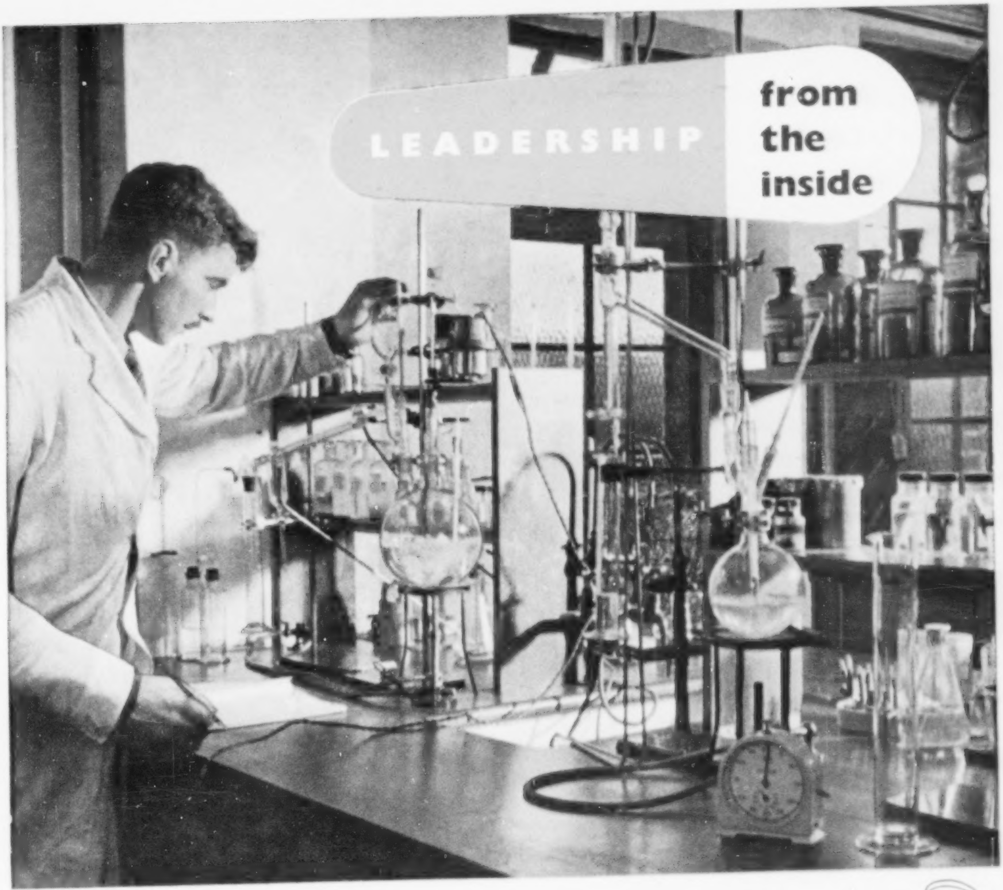

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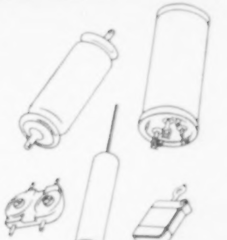

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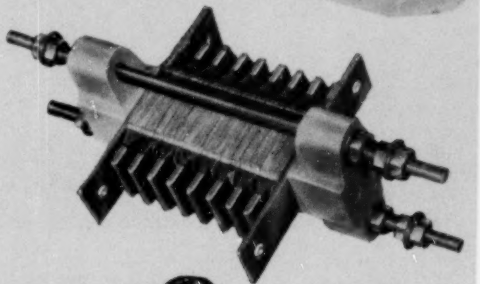

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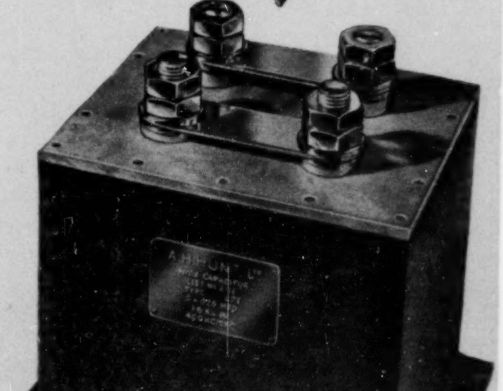

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